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FINAL TECHNICAL REPORT

BROADBAND ANTENNA MEASUREMENT TECHNIQUES

by
J. D. Adams, F. L. Cain, and C. E. Ryan Jr.

Radar Division
Engineering Experiment Station
Georgia Institute of Technology

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FOREWORD

This final technical report describes the work performed under Contract F30602-73-C-0914, Project No. 4506, Task No. 450604 during the period from 16 February 1973 to 19 March 1974.

The work on this contract was performed by the Radar Division of the Engineering Experiment Station at Georgia Institute of Technology, Atlanta, Georgia under Project A-1517, with the support of Scientific-Atlanta, Inc., under sub-contract. The RADC Project Engineer was Mr. Martin Jaeger (OCTS).

In addition to the authors, contributors to the work described herein include D. C. Griffin, T. M. Miller, Jr., and J. M. Schuchardt at Georgia Tech, J. S. Hollis and R. E. Pidgeon, Jr. at Scientific-Atlanta, Inc.

This technical report has been reviewed and it is approved.

Approved:

Martin Jaeger
Project Engineer
Radiation and Signal Processing Section

ABSTRACT

This report presents the results of a program to study and investigate advanced measurement techniques for evaluating the performance of broadband antenna systems for use in high resolution radar systems. New techniques to measure, record, and analyze antenna gain and pattern performance were studied, and the feasibility of developing the necessary instrumentation to perform these measurements was investigated. Systems based on the use of broadband noise signal sources and systems using sweep frequency techniques were studied. It was concluded that systems using broadband noise signal sources would not be cost-effective. Preliminary design of a broadband amplitude-only electronically integrating sweep frequency system was completed. It was concluded that this system could be implemented immediately and that it would provide an effective first step in realization of the ultimate phase plus amplitude broadband antenna measurement system. The implementation and operation of sweep frequency amplitude plus phase systems were studied, and an effective approach to the realization of a phase/amplitude system was identified. Recommendations for implementing this approach are presented.

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SECTION I

INTRODUCTION

A. GENERAL

New antenna-range measurement techniques must be developed to characterize antennas that are designed to meet the requirements of modern-day high-resolution radars. Satisfying the mission objectives of planned or existing modern-day radars requires meeting stringent performance specifications on, among other things, range resolution and range accuracy. Range resolution of 6 inches or less is desired in some applications. Obtaining this type of range resolution, along with good range accuracy, requires a very narrow pulse with a fast rise time to provide a well defined leading-edge. Obtaining increased resolution and accuracy requires increasing the bandwidth of several stages of the overall radar system, including that of the antenna. The old rule of thumb that a 1-microsecond pulsewidth, which requires a 1-MHz bandwidth, will provide a 500-foot resolution must now be extended to a 1-nanosecond pulsewidth, which requires a 1-GHz bandwidth, provides a 6-inch resolution. In order to obtain a one-half inch resolution, as apparently some users are proposing, a one-tenth nanosecond pulsewidth and a bandwidth of 10 GHz will be required. To radiate these broadband pulses with essentially no distortion, the antenna must have a wide instantaneous bandwidth.

Because short-pulse directive antennas must have a wide instantaneous bandwidth, new techniques for quickly measuring and evaluating antenna performance must be developed. Not only must the antenna have good broadband spatial amplitude characteristics, but it must also have good phase properties over the broad bandwidths required. Antenna phase and/or amplitude errors can lead to such undesirable effects such as degraded spatial patterns, pulse distortion, and limitations on pulse compression performance. In many cases, if the specific phase and amplitude errors were known, corrective measures could be implemented to achieve more nearly ideal performance. Thus, it becomes imperative that the instantaneous broadband phase and amplitude properties of the antenna be known.

The measurement of instantaneous broadband phase and amplitude properties of an antenna to determine its effects on short pulses is a deviation from the conventional antenna pattern measurement technique. The additional measurement requirements that must be met to characterize short-pulse radar antennas demand a significant increase in antenna measurement capability that does not now exist. Currently available broadband measurement techniques which are applicable to high gain radar antennas typically rely on radio stars or noise generators as signal sources; as a consequence, they suffer such deficiencies as lack of practical bandwidth control, lack of phase information, limited dynamic range, and usually no polarization discrimination. Although time-domain impulse response techniques have also been used for testing wideband antennas [1,2], these techniques are generally not well suited to the high-gain narrow-beam microwave antennas. Under this program, Georgia Tech has investigated new techniques for achieving the desired broadband phase and amplitude measurement capabilities.

B. PROGRAM OBJECTIVES

The general objectives of this contract are to explore new techniques to measure, record, and analyze broadband antenna performance and to determine the feasibility of developing necessary instrumentation to perform such measurements. Although emphasis has been given to the 1-15 GHz frequency range, availability of major instrumention components for the 200 MHz to 75 GHz region was investigated.

As a minimum in achieving the objectives of this program two basic concepts, which are designated as Concepts I and II, were considered as well as any variations from these two basic concepts. The first concept involves the use of a broadband leveled noise source to provide average antenna gain and pattern information. The second concept involves a swept frequency transmitter and a suitable receiver for measuring and recording sampled phase and amplitude data over a broad bandwidth. The recorded phase and amplitude data would permit the investigation of antenna response to any particular frequency component or to any combination of frequency components; hence, by proper selection of frequencies, the antenna effect on pulses of various bandwidths and modulations can be simulated and diagnosed. A variation from Concepts I and II, designated as the Hybrid approach, was formulated and investigated. This Hybrid approach has some similarities to both Concepts I and II. In all approaches, the system sensitivity should be sufficient to permit system use on very long outdoor antenna ranges with a dynamic range of at least 40 dB. A transmitter power output capability of approximately 1 to 10 watts CW is desired.

C. SUMMARY OF WORK PERFORMED

Initially, efforts on this contract were devoted to the analysis and evaluation of instrumentation for the wideband measurement systems of Concepts I and II. Of the various Concept I systems which were considered, two of the most promising ones were selected for evaluation. These two were identified as (1) a Broadband Amplified Thermal Noise System, and (2) a Noise-Modulated Synthesized Broadband Noise System. Under Concept II, both computer controlled amplitude-only systems and computer controlled amplitude-plus-phase systems were evaluated. A third approach, designated as the Hybrid system because of its similarities to both Concept I and Concept II systems, was also synthesized and evaluated. Relative evaluations among all of the Concepts I and II systems indicated that the Concept I systems would not be cost effective due to their measurement limitations and equipment complexities. Consequently, with sponsor concurrence, succeeding efforts were concentrated on completing design of the Hybrid system and on further evaluation of techniques for implementing a Concept II system. A brief summary of the results of the investigations in all of the various areas is given below.

1. Concept I--Broadband Noise

From the various approaches that were considered for implementing a Concept I system, two different measurement systems involving noise techniques were selected for evaluation. One of these two noise techniques involves the use of a very broadband noise source, octave bandwidth TWT power amplifiers, a broadband transmitting antenna, and a broadband receiver whose pre-detection bandwidth is as broad as the transmitted spectrum. A simplified block diagram of the transmitter for this system is shown in Figure 1. Continuous center frequency control from 1 to 15 GHz and bandwidth adjustment from nearly 0 to 1 GHz are achieved through heterodyne techniques with adjustable frequency local oscillators and fixed-tuned filters. The most appropriate receiver type for this application is the wideband RF which consists of a detector-video receiver proceeded by a low-noise RF amplifier. A block diagram of this receiver is shown in Figure 2. This receiver has adequate sensitivity and bandwidth, but a separate RF amplifier is required for each octave of frequency or waveguide frequency band.

The second noise technique is based on sequential transmission of band-limited frequency-contiguous noise which is averaged in the receiver to synthesize a very broadband noise response. A block diagram of this system is shown in Figure 3. This system consists basically of a conventional voltage

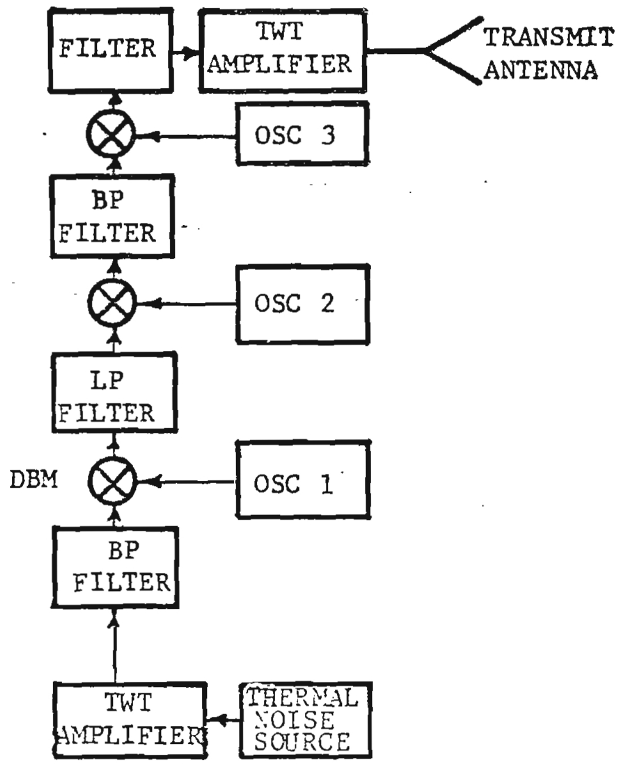


Figure 1. Simplified block diagram of thermal noise transmitter

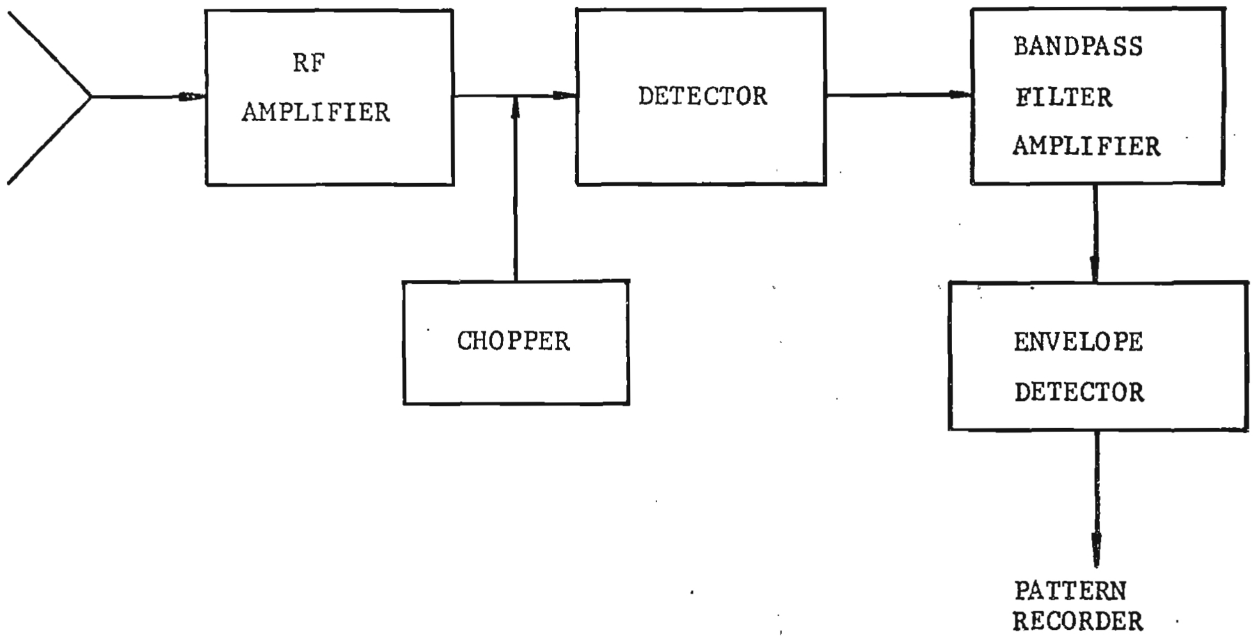


Figure 2. Broadband receiver for thermal noise transmitter

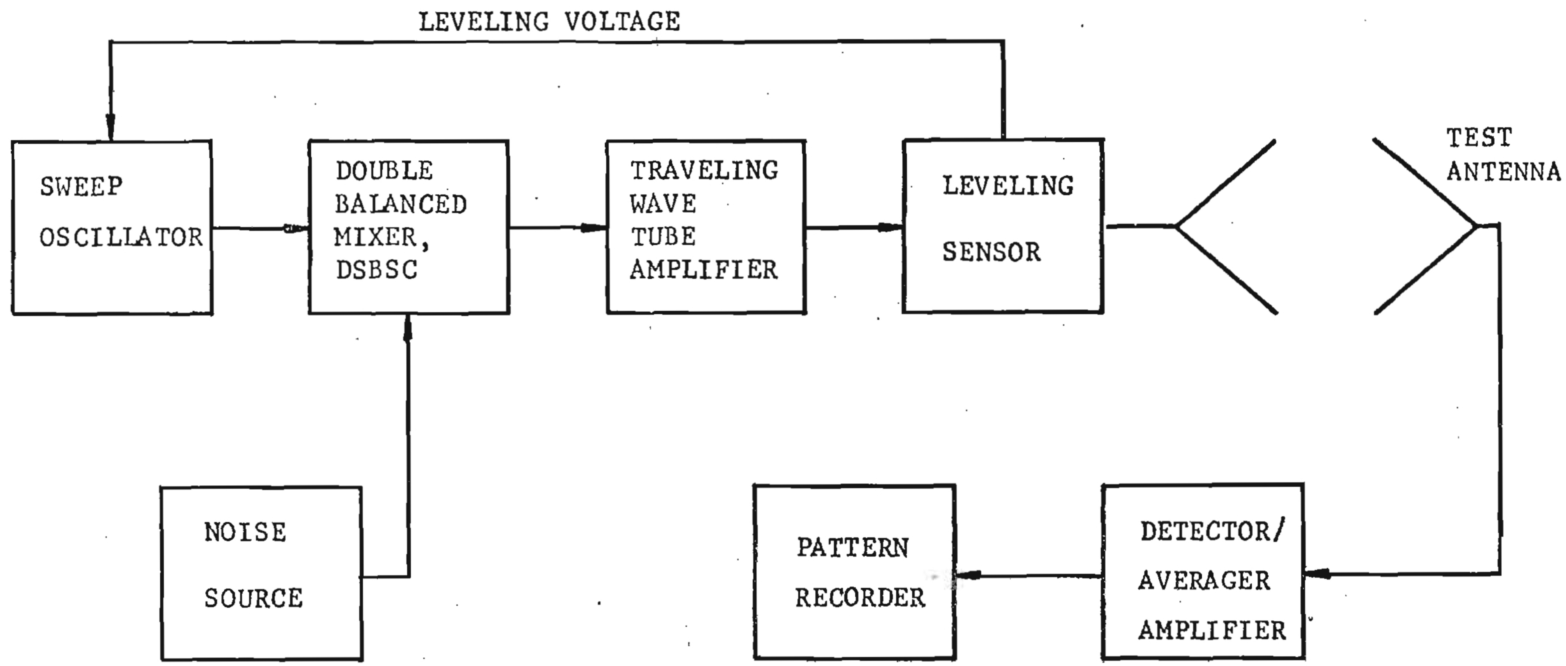


Figure 3. Noise-modulated synthesized broadband system

controlled oscillator, noise generator, and TWT power amplifier at the transmitting site with an ocatave bandwidth receiver and averager at the receiving site. Operation of this system may be understood by reference to Figure 4. The antenna response to each of the relatively narrow-band noise pulses is determined by the receiver, and this response is integrated over the time required to step the noise pulses over the full bandwidth desired. Consequently, an adjustable bandwidth capability is achieved.

For the received power to reflect only the test antenna response over the selected bandwidth, a constant power density versus frequency is required at the test antenna. With a broadband noise spectrum, satisfying this requirement is complicated by the fact that transmit antenna gain and space loss both depend on frequency. Thus, even if a uniform amplitude noise spectrum at the required power level is generated, the required constant power density at the test antenna may not be achieved. One possible approach to a solution of this problem is to tailor antenna gain to vary inversely as space loss so that these two effects cancel. However, tailoring antenna gain to accurately follow a prescribed gain function over an octave bandwidth is a difficult task. The Concept I broadband noise systems would provide average amplitude information only, and a new spatial pattern must be measured for each desired bandwidth.

2. Concept II--Phase and Amplitude System

The function of the phase-amplitude system is to measure and digitally record the phase and amplitude response of the test antenna as the transmitter is swept over a wide frequency range (typically up to 1 octave). These phase and amplitude data as a function of frequency can be digitally processed to determine antenna response to wideband signals, such as various short pulses or FM coded longer pulses. Two different phase measurement techniques have

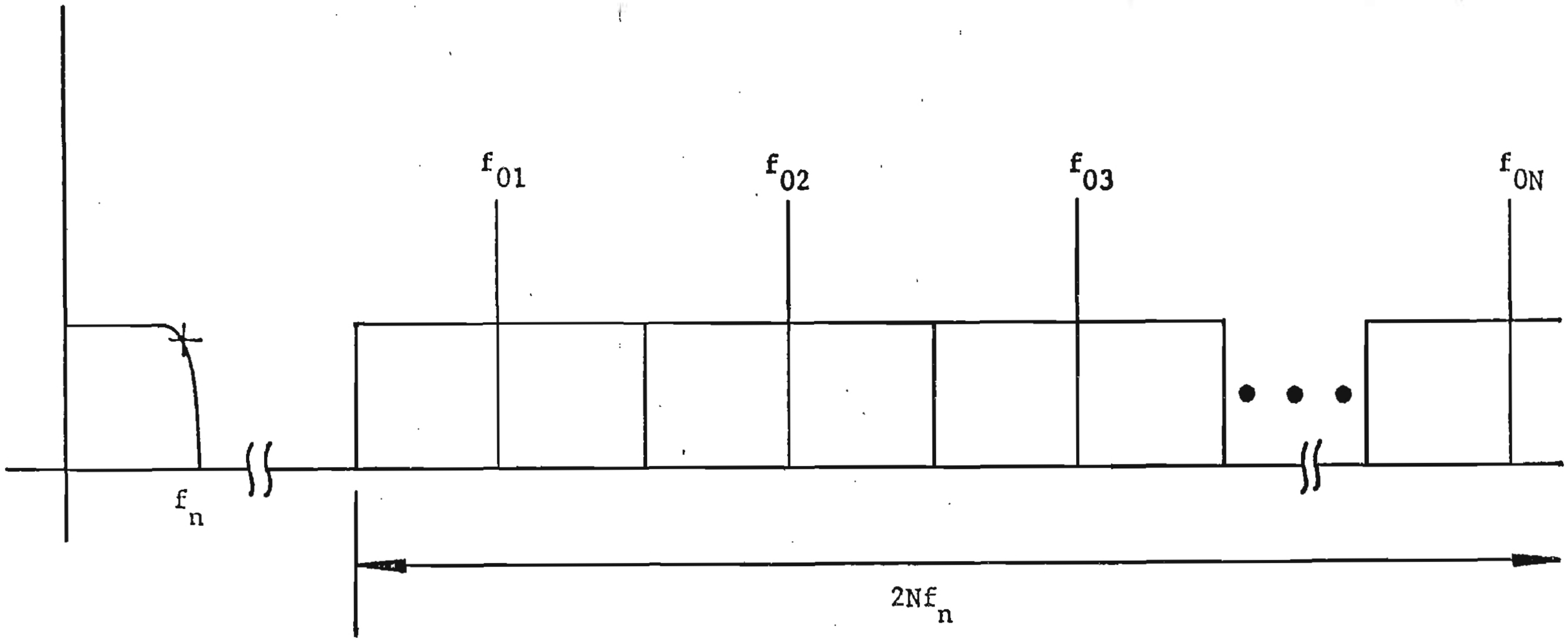
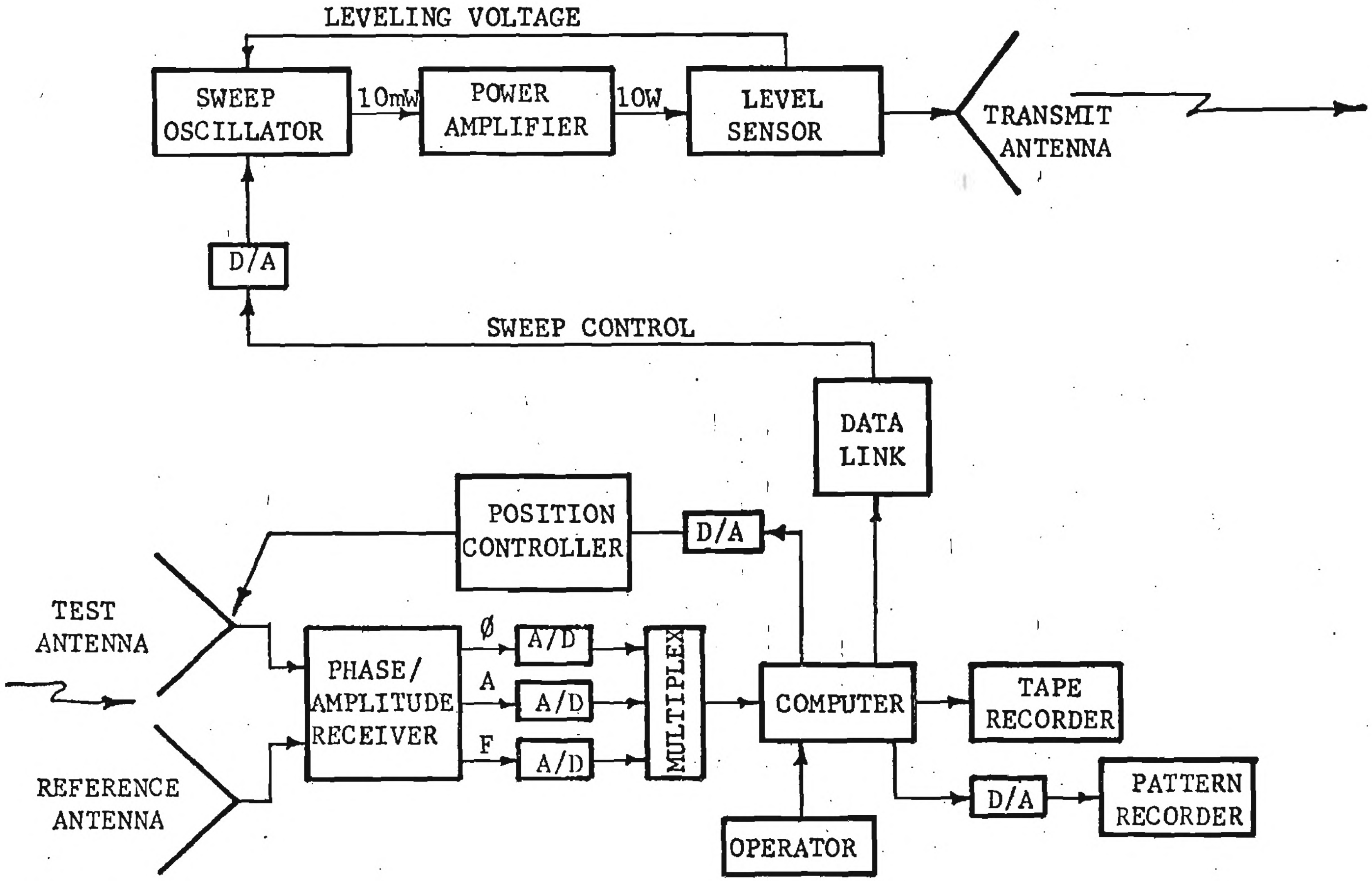


Figure 4. Operation of synthesized broadband system

been investigated for this system. With one technique, the transmitter produces a single-tone signal which is swept across the desired frequency range. With this technique, phase delay is measured directly, and a separate RF reference channel is required to obtain a phase reference signal at the receiver. With the second technique, the sweep oscillator is modulated to produce a carrier plus an upper and a lower sideband. These three signal components are swept across the measurement range while the sideband separation remains constant. With this second technique, the receiver separates the two sidebands whose phase separation has been modified after passing through the test antenna and then measures the phase between them. From this sideband phase difference as a function of carrier frequency, the test antenna phase characteristics can be obtained by integration. This second technique of measuring sideband phase information is based on the concept of group velocity, i.e., measurement of group delay. A phase/amplitude system using the direct (separate reference channel) phase measurement technique is diagrammed in Figure 5, while a system using the group velocity technique is diagrammed in Figure 6.

The two basic differences between these phase/amplitude systems are that (1) the group velocity system requires the addition of a modulator in the transmitter, and (2) the group velocity system requires no separate calibrated reference channel. A more subtle difference between the two systems involves the procedures required for system calibration and for data reduction. With the group velocity system, the measured sideband differential phase data will depend, of course, on the phase properties of the entire transmission path from signal generation to detection. To obtain test antenna phase properties, the measured data must be corrected for any phase dispersion introduced by the transmitter or the transmit antenna and for any errors in the receiver. The



Ø = PHASE
A = AMPLITUDE
F = FREQUENCY
A/D = ANALOG-TO-DIGITAL
D/A = DIGITAL-TO-ANALOG

Figure 5. Direct phase/amplitude separate reference channel system

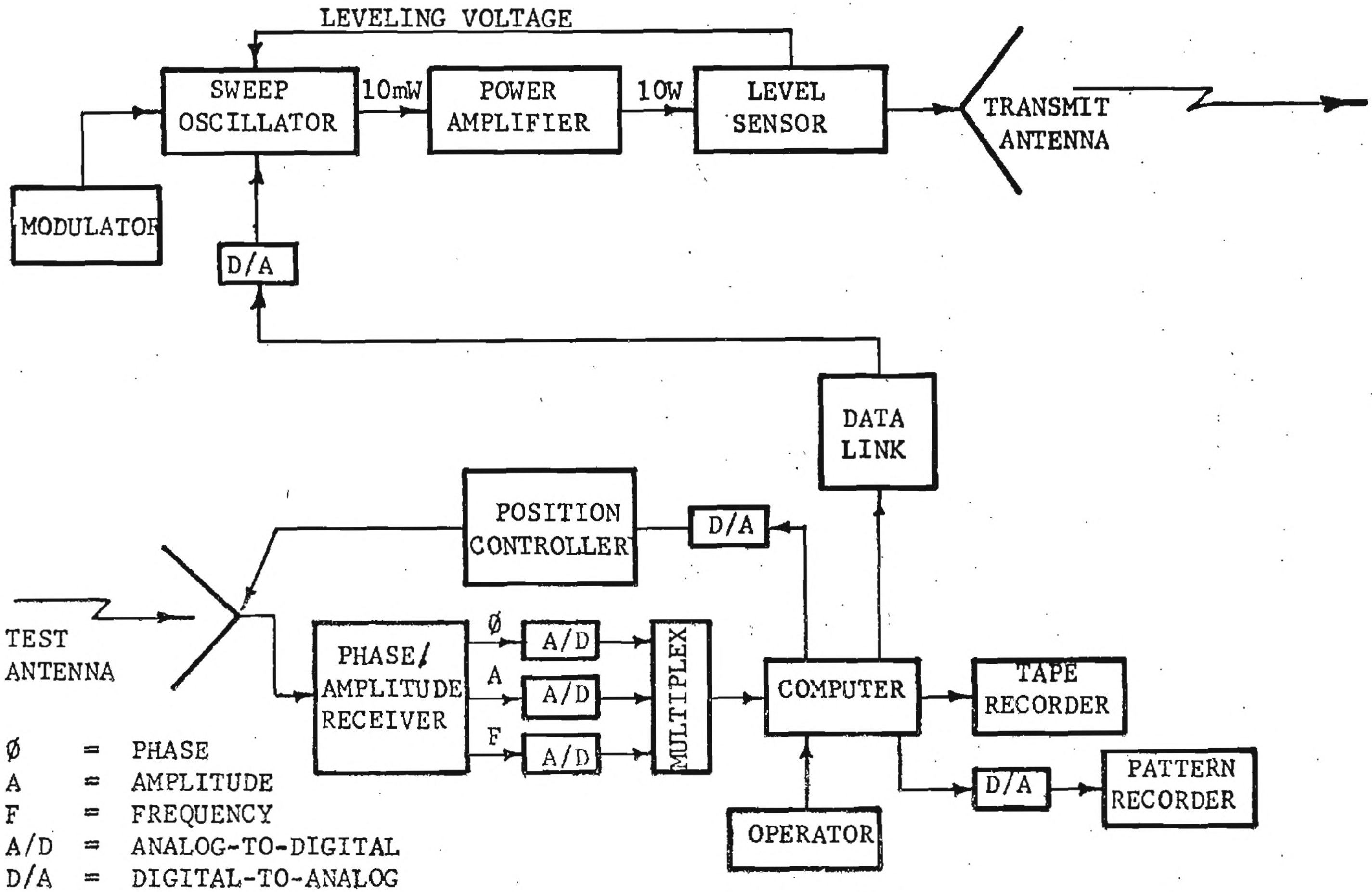


Figure 6. Group velocity concept phase/amplitude system

measured data can be computer corrected if calibration data versus frequency is determined and stored for this purpose. With the separate reference channel system, the calibration involves determining the phase properties of the reference antenna and determining any difference in phase properties of the two transmission paths from the receiver to the test antenna and from the receiver to the reference antenna. Again, calibration data versus frequency can be used to computer-correct the measured data. Techniques for calibration and operation of these systems are discussed in Section IV.

3. Hybrid System

As indicated previously, measurement flexibility of a noise-based system is limited, and in addition, the instrumentation for a uniform amplitude adjustable bandwidth, noise transmitter is quite complex. A system which provides increased measurement capability with decreased instrumentation complexity is illustrated by the simplified block diagram shown in Figure 7. This new system uses a sweep frequency transmitter, such as would be used for the phase/amplitude system, and it provides results that are equivalent to the broadband noise systems. This system can also provide CW amplitude-only single frequency data. Thus, it is identified as the Hybrid system.

Operation of the Hybrid system may be easily understood by consideration of how an average antenna response is obtained with a noise system. The receiv output for the noise system is proportional to the integral of the amplitude transfer function of the antenna over the frequency range of the noise source. Consequently, the integration process of the noise system is instantaneous by virtue of the fact that all of the frequency components are simultaneously present in the incident signal and that a power detector provides a voltage which is proportional to the total power of all of these frequency components of the incident signal. The Hybrid system produces "average" antenna gain,

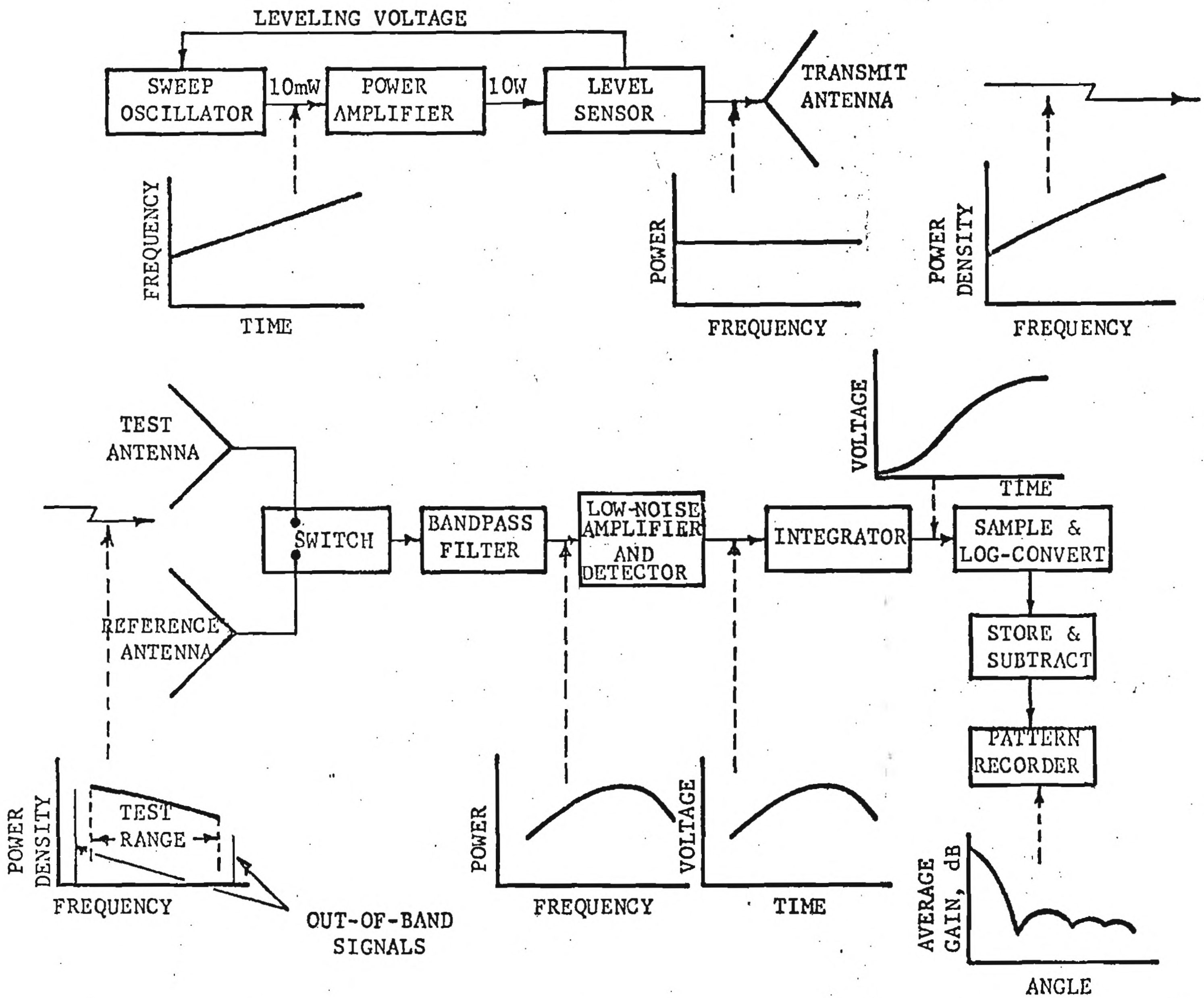


Figure 7. Simplified block diagram of hybrid system

which is equivalent to that of the noise system, by integrating the response of the antenna to a signal which is swept over the desired frequency measurement range. That is, in the Hybrid system, time-domain integration of a swept frequency signal replaces frequency integration of a noise signal.

Important waveform considerations at various circuit points in the Hybrid system are depicted by the sketches with dotted lines indicating the appropriate circuit point on the block diagram in Figure 7. An important feature of this Hybrid system is that amplitude variations (such as those due to variation in transmit antenna gain and space loss) can be compensated automatically in the receiver. This compensation is equivalent to obtaining constant power density at the test antenna aperture. Basically, compensation is based on electronically determining the ratio of test antenna response to the response of a reference antenna whose gain versus frequency characteristics are known. This system can provide average antenna response over any bandwidth up to the transmitter sweep range, or if desirable, the transmitter can be operated in a nonsweep mode so that the antenna response at any selected single frequency can be recorded. Preliminary design of this system is complete, and it is essentially ready for implementation.

All aspects of these systems, such as hardware requirements and system operation, are described in detail in following sections of this report. The body of this report describes the Concept I systems which were evaluated. Design and operation of the hybrid system is discussed, and the results of the feasibility studies of phase/amplitude systems are presented. To preserve report continuity and to facilitate reading of the report, basic component descriptions are presented in Appendix I.

SECTION II

CONCEPT I SYSTEMS

A. INTRODUCTION

Initial efforts on this contract were devoted to investigation and evaluation of both Concept I and Concept II systems. Several Concept I systems were postulated, and two were investigated to the detail required to evaluate their potential as a broadband antenna measurement system. A hardware description of these two systems is given in this section, along with a discussion of the factors which leads to a conclusion that a Concept I system is not a cost-effective approach to the broadband antenna measurements problem.

B. AMPLIFIED THERMAL NOISE SYSTEM

A block diagram of the amplified thermal noise transmitter is shown in Figure 8. This transmitter may be considered as being composed of two sections One section provides for adjustable bandwidth, and the other section provides adjustable center frequency. The particular frequency range of the devices has been selected for ease of implementation. This transmitter is based on use of a broadband uniform noise source, a low-noise TWT, and a broadband TWT amplifie as described in Appendix I, along with fixed tuned filters and heterodyne techniques to achieve an adjustable center frequency and bandwidth. Also shown in Figure 8 is a frequency scale upon which the bandpass or spectral output chrracteristics of the various components are shown. These simplified characteristics are identified with the respective components by connecting lines. For discussion, the bandpass and spectral output characteristics have been idealized and integer frequencies are used.

The thermal noise is amplified first by a low-noise TWT amplifier and then filtered by a bandpass filter such that the spectrum into the first mixer

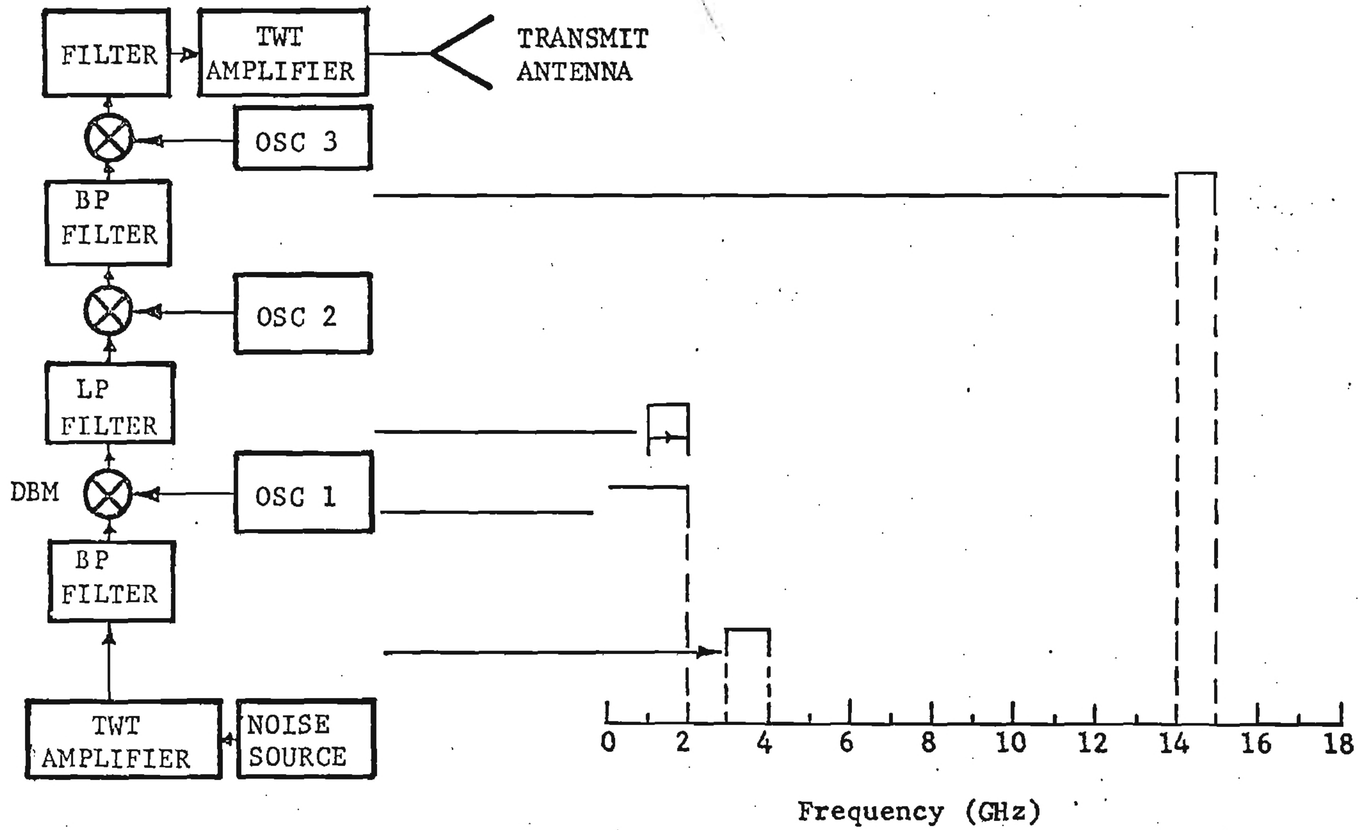


Figure 8. Amplified thermal noise transmitter

has a power level of about 0 dBm (which is generally required for good mixer performance) and a constant amplitude from 3 to 4 GHz. Because the purpose of this stage of the system is only to provide an adjustable bandwidth, the frequency as well as the frequency range has been chosen to simplify mixer and local oscillator requirements. The output from oscillator 1 is mixed with this spectrum in a double-balanced mixer. The frequency of oscillator 1 is tunable from 5 GHz to 6 GHz. The double-balanced mixer suppresses the two input signals (original noise spectrum and oscillator 1 signal) relative to the sum and difference of these two signals. The first mixer output spectrum consists of a lower and an upper sideband. A low pass filter having an upper cut-off frequency of 2 GHz is used to pass the lower sideband and reject the upper. If oscillator 1 is set at 5 GHz, then the lower or difference sideband spectrum will cover from 1 GHz to 2 GHz. Now, if the frequency of oscillator 1 is increased, the first mixer output spectrum will shift up in frequency and the lower sideband will be partially outside the passband of the low pass filter. By increasing the frequency of oscillator 1 to 6 GHz, the lower sideband will be completely outside the filter passband. As shown by the notation on the frequency scale, the upper limit of the noise spectrum at the output of the low pass filter remains at 2 GHz and the lower limit is adjustable from 1 to 2 GHz. Thus, the bandwidth is adjustable from 0 to 1 GHz.

The next stage of the transmitter is for shifting the adjustable bandwidth spectrum to other portions of the 1-15 GHz frequency range of interest. To accomplish this shift, the adjustable bandwidth spectrum is applied to a second mixer. This mixer is used as a double-balanced modulator where the spectrum is up-converted by mixing with a 16 GHz signal from the fixed-frequency oscillator 2. The output spectrum is a double sideband suppressed carrier signal, and only the lower sideband is passed by a bandpass filter. At the output of this bandpass filter, the spectrum bandwidth is adjustable from 0 to 1 GHz across the frequency range from 14 to 15 GHz. The third mixer and adjustable oscillator 3 are used to down-convert this spectrum to the desired center frequency. By adjusting oscillator 3 output from 16 to 30 GHz, the lower sideband from mixer 3 can be translated to any center frequency from 1 to 15 GHz. The upper sideband and the oscillator 3 signals are rejected by a bandpass filter. The final amplifier (or amplifier chain) must have sufficient gain (50 to 60 dB) to amplify the noise spectrum to a level of 1 to 10 watts.

Because microwave components, especially amplifiers, typically operate over octave or waveguide bandwidths, coverage of the frequency range of interest (1-15 GHz) would require that the third mixer, the third oscillator, the final amplifier, and the bandpass filter be duplicated for each octave or waveguide band. Several other components, not shown in the simplified block diagram of Figure 8, would be required for proper system operation. To obtain equal noise power across the frequency bands, equalization of TWT gain variations would be required. To prevent reactive mismatches between components, an isolator would be required before each filter to absorb reflected power.

All oscillators in the system must be stabilized to a low percentage of error because the spectrum bandwidth and the center frequency are controlled by the frequency differences between the three relatively high frequency oscillators. A large collection of microwave components is thus required to build this transmitter. A broadband transmitting antenna is required to radiate this signal. A transmitting antenna gain function which will provide constant power density at the test antenna is desirable.

A suitable broadband receiver is necessary to record the average radiation pattern of an antenna over a broadband of frequencies. Assume for the moment that the broadband noise signal arrives at the test antenna aperture with constant spectral density, i.e., it has constant power per unit bandwidth. If such a signal in incident on an antenna, the output spectral density is determined by the amplitude-only frequency response characteristics, H(s), of the antenna. Broadband detection of this output signal followed by low-pass filtering with a filter having an integration time sufficient to remove the random amplitude fluctuations of the noise source results in an output that is proportional to $\int H(s)\,ds$. As indicated in Appendix I, the most suitable receiver type is the broadband RF receiver. This type of receiver, without the low-noise amplifier, is used, for example, in the HP 415 SWR Meter and the SA 1554-2 Crystal-Bolometer Amplifier plug-in for the SA 1520 Rectangular Pattern Recorder.

C. NOISE-MODULATED SYNTHESIZED BROADBAND SYSTEM

A simplified block diagram for implementation of this type of system was shown in Figure 3. The system would consist basically of a conventional voltage controlled oscillator (e.g., a BWO), noise generator, double-balanced mixer, power amplifier (TWT), and amplitude leveler at the transmit site with octave bandwidth receiver (RF front end and detector) at the receiving site. The wideband antenna response would be synthesized by integrating response to narrow band noise modulated signals.

The low level portion of the noise transmitter considered here consists of a voltage-tunable oscillator, a double balanced mixer, and a low frequency noise source. The operation of the low-level portion of the noise source is most influenced by the effectiveness of the double balanced mixer (DBM) as a Double Sideband Suppressed Carrier up-converter. If the DBM functions effectively, the oscillator signal is suppressed, and the low frequency noise is up-converted to produce well characterized noise sidebands centered about the oscillator frequency.

Figure 9 shows the spectrum of the signals involved. A low frequency spectrum extending from DC to f_n is generated by the noise source. This low frequency noise is translated upward and is centered around the CW oscillator frequency f_o. The actual CW carrier is suppressed by the double balanced mixer. This double sideband suppressed carrier modulation (DSBSC) in effect doubles the noise spectrum bandwidth.

To cover the desired antenna measurement range, the voltage tuned oscillator signal is slowly swept over the required frequency range, and the broadband noise response is determined electronically at the receiver. This procedure was illustrated in Figure 4.

D. INTERFERENCE BY NOISE SYSTEM

A potential problem with use of a broadband noise system is interference by the radiated signal with other receivers in the surrounding area. Because the radiated signal could be of relatively high power density (10 watts radiated by high gain antenna) and very broadbanded, interference could be significant. Calculations were made to determine the expected received noise power for various receiver bandwidths and center frequencies.

The noise power per unit bandwidth, N_d, at the receiving antenna terminals is given by

$$N_d\ (\varphi, \theta) = \frac{P_t\ G_t\ (\theta)\ G_r\ (\varphi)\ \lambda_c^2}{(4\pi)^2\ R^2\ B_t} \qquad (1)$$

where

P_t = power transmitted,

$G_t\ (\theta)$ = power gain of transmitting antennas as a function of angle θ,

$G_r\ (\varphi)$ = power gain of receiving antenna as function of angle φ,

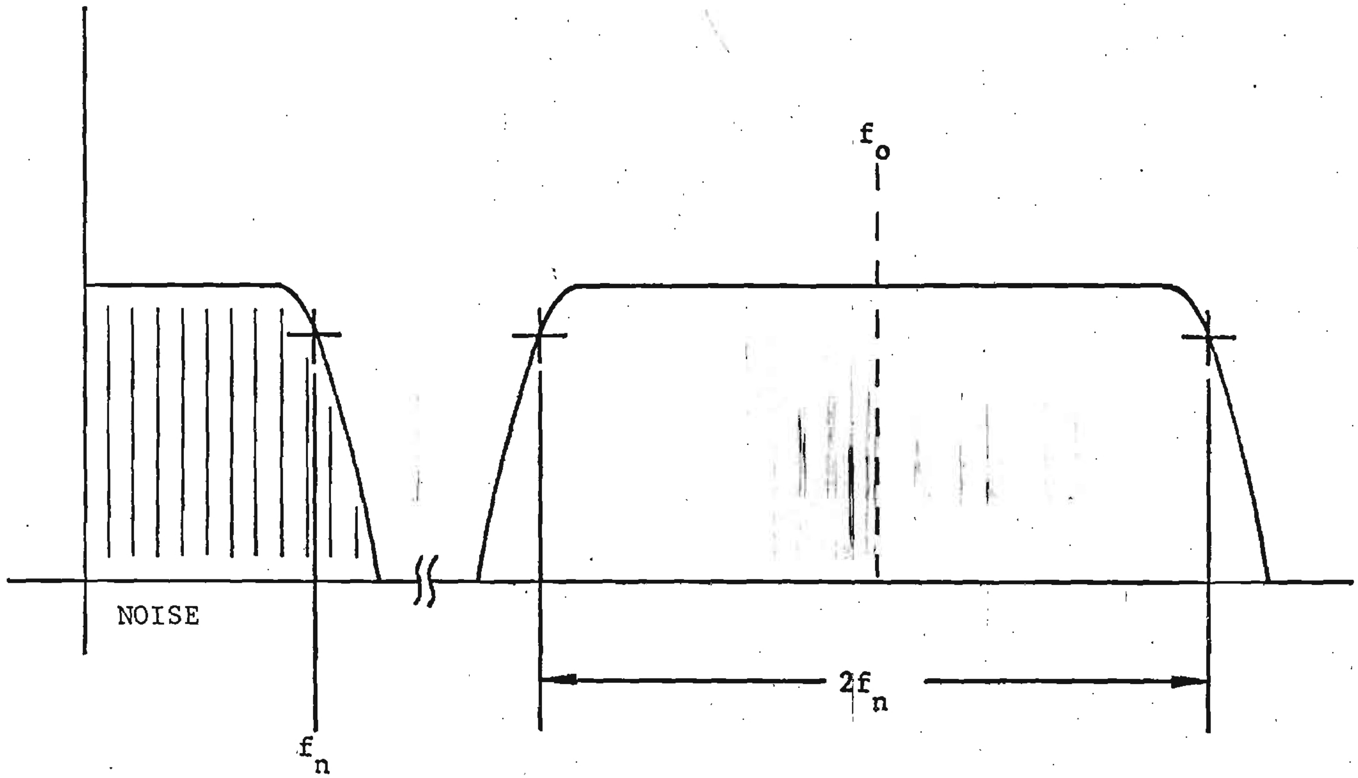


Figure 9. Double-balanced mixer response

λ_c = wavelength at the carrier frequency,

R = range between transmitting and receiving antennas, and

B_t = noise bandwidth.

The geometry of the interference situation is shown schematically in Figure 10. For analysis, the above parameters were assumed to have values typical of those to be expected under measurement conditions. Noise bandwidths of 250 MHz and 1 GHz at carrier frequencies between 0.5 and 14.5 GHz and a transmitter power of +15 dBm were assumed. Any variations of the transmitting and receiving antenna gains over the noise bandwidths were neglected. Although several different transmitting antennas could be used over the 1-15 GHz range, the peak gains of the transmitting and receiving antennas were assumed to be equal and to have a value of 20 dB up to 8 GHz and of 30 dB from 8 GHz to 15 GHz. In summary, the following parameters were assumed:

B_t = 250 MHz and 1 GHz

P_t = +15 dBm, and

$$G_t(0) = G_r(0) = \begin{cases} 20 \text{ dB @} f < 8 \text{ GHz or} \\ 30 \text{ dB @} f \geq 8 \text{ GHz.} \end{cases}$$

Calculations were made for several different center frequencies, and the results are given in Table 1. These data show the noise power density at the receiving antenna terminals for main lobe-to-main lobe coupling with the assumed maximum gains. Sidelobe coupling can be accounted for by subtracting the sidelobe level in dB from the value of N_d which is presented. The receiver bandwidth must be accounted for as follows. If the bandwidth of the receiver up through the detector is greater than or equal to the noise bandwidth B_t, the full noise power will be detected and would be equal to the noise power per unit bandwidth N_d multiplied by the ratio of B_t to 1 MHz. For a superheterodyne re-

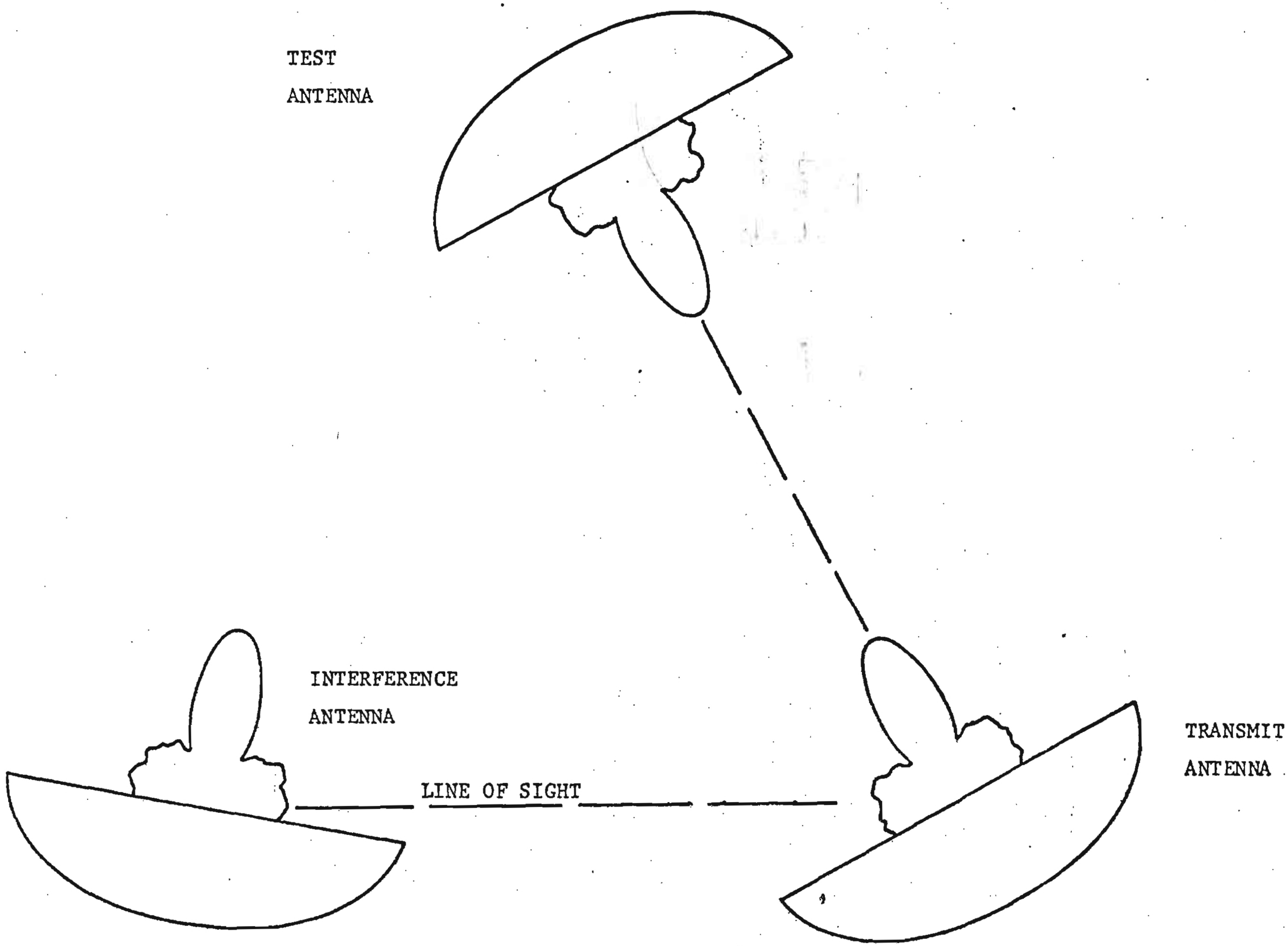


Figure 10. Schematic of interference condition

TABLE 1

RECEIVED NOISE POWER DENSITY

$f_c = 0.5$ GHz		$f_c = 0.5$ GHz		$f_c = 5$ GHz		$f_c = 5$ GHz	
$B_t = 250$ MHz		$B_t = 1.0$ GHz		$B_t = 250$ MHz		$B_t = 1$ GHz	
$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)
-45	1.0	-51	1.0	-65	1.0	-71	1.0
-51	2.0	-57	2.0	-71	2.0	-77	2.0
-57	4.0	-63	4.0	-77	4.0	-83	4.0
-65	10.0	-71	10.0	-85	10.0	-91	10.0
-71	20.0	-77	20.0	-91	20.0	-97	20.0

TABLE (Concluded)

RECEIVED NOISE POWER DENSITY

f_c = 10 GHz		f_c = 10 GHz		f_c = 14.5 GHz		f_c = 14.5 GHz	
B_t = 250 MHz		B_t = 1 GHz		B_t = 250 MHz		B_t = 1.0 GHz	
$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)	$N_d\left(\frac{dBm}{MHz}\right)$	R(km)
-51	1.0	-57	1.0	-55	1.0	-61	1.0
-57	2.0	-63	2.0	-61	2.0	-67	2.0
-63	4.0	-69	4.0	-67	4.0	-73	4.0
-71	10.0	-77	10.0	-75	10.0	-81	10.0
-77	20.0	-83	20.0	-81	20.0	-87	20.0

ceiver with an IF bandwidth which is less than B_t, the detected noise is the indicated noise power density multiplied by the ratio of the IF bandwidth to 1 MHz.

Several conclusions can be made with the aid of the data which are presented in Table 1. If recording an antenna pattern in conjunction with a broadband receiver, the total noise power received would be N_d increased by 14 dB for the 250 MHz case and by 30 dB for the 1 GHz case. For pattern recording under either the 250 MHz or the 1 GHz condition at ranges of 10 km (6.2 miles) or less, the received power under the assumed conditions is adequate for observing at least -40 dB sidelobes with a typical receiver sensitivi For interference considerations without main-lobe-to-main-lobe coupling, N_d would decrease greatly for most high gain antennas (25 to 50 dB depending on the orientation of the receiving antenna with respect to the transmitting antenna). If the victim receiver were a typical superheterodyne with a 10 MHz IF bandwidth, then the detected power at a range of 4 km for the f_c = 0.5 GHz B_t = 250 MHz case would be -57 + 10 - 25 = -72 dBm and -57 + 10 - 50 = -97 dBm for antenna coupling reduced by 25 and 50 dB, respectively. At range of 20 km these levels decrease to -86 dBm and -111 dBm, respectively. Thus, depending on the orientation and the gain of the antenna at the victim reciever site, interference could very well be a problem. A specific analysis would be required for each anticipated test range to determine permissible transmitted power levels.

E. SUMMARY AND EVALUATION OF CONCEPT I SYSTEMS

Concurrent with the investigation and evaluation of the broadband noise systems which have been described in this section, swept frequency phase and/or amplitude systems were also under investigation. These concurrent investigations for realization of broadband antenna measurement instrumentation were required in order to define the relative merit of the two basic concepts under consideration. Based on these initial investigations, the following conclusio and evaluations of Concept I systems are made.

(1) Because the noise spectrum across the entire measurement bandwidth i simultaneously present, spectrum shaping must be accomplished by passive devices, and therefore, achieving uniform power density at the test antenna would be difficult.

(2) Only average antenna patterns can be obtained, and a new spatial pattern must be recorded for each desired bandwidth.

(3) Achieving adjustable bandwidth, adjustable center frequency, and adequate transmitter power requires a large collection of microwave equipment.

(4) Because radiation at potentially interfering frequencies cannot be easily controlled, a broadband high power noise transmitter could cause significant interference.

On the other hand, a Concept II system or the Hybrid system(Sections IV and III, respectively) can provide average patterns at selected bandwidths and center frequencies by simply changing sweep limits at the transmitter. These other systems can also provide single frequency data at any frequency within the transmitter sweep range. In addition, these other systems (as discussed later) do not require uniform power density at the test antenna. Moreover, implementation of the Hybrid system, which yields results equivalent to the noise systems, is considerably less complex than implementation of the noise systems. Further, because recorded broadband phase and amplitude data would provide complete flexibility in antenna diagnostics (CW spatial patterns, avera patterns, pulse diagnostics, fm distortion prediction), development of a phase/

amplitude computer controlled system is the ultimate goal of the broadband antenna measurements program. Realization of the Hybrid system will be a positive step in the realization of the phase/amplitude system, while at the same time, it will provide more capability than a noise-based system could provide. Based on the conclusion that a more detailed design analyses of Concept I systems was not appropriate, efforts were thereafter devoted to only the Hybrid and Concept II systems. The remainder of this report deals with the design, capabilities, and operation of these other systems.

SECTION III

HYBRID SYSTEM

A. INTRODUCTION

As indicated in the previous section, Concept I systems provide only limited measurement capabilities while having complex instrumentation requirements. The Hybrid system, which is less complex than the noise systems of Concept I, can provide both the average pattern measurement capability of a Concept I system and the single frequency capability of an amplitude-only Concept II system. Measurement of broadband spatial patterns can be considered as a first step in the characterization of broadband radar antennas. If the antenna under test does not have a well behaved broadband pattern, it may not provide satisfactory radar performance, regardless of its phase properties which affect the pulse characteristics. In addition, broadband spatial characterization of an antenna is important from the electromagnetic compatibility (EMC) and electronic countermeasure (ECM) viewpoints. Thus, measurement of average broadband spatial patterns is an important consideration. The Hybrid system will provide this capability, and because of its relative simplicity, the Hybrid system can be configured as a mobile system for field test. Thus, the Hybrid system will provide an important measurement capability while serving as a building block for the phase/amplitude system.

A simplified block diagram of the Hybrid system has been given in Figure 7. The Hybrid system produces "average" antenna gain by integrating the antenna response to a signal which is swept over the desired frequency measurement range. That is, in the Hybrid system, time-domain integration of a swept frequency signal replaces frequency integration of a noise signal. An important feature of this Hybrid system is that amplitude variations (such as those due to variations in transmit antenna gain and space loss) can be compensated automatically in the receiver. However, because a constant-power transmitter may be desirable for some applications and because this transmitter might be used with other receivers, this design includes provisions for obtaining a constant-power transmitter.

Referring to Figure 7, operation of the hybrid system may be described as follows. The reference antenna switch position is selected and the sweep oscillator begins to sweep over a pre-selected frequency range. The power amplifier (a travelling wave tube) provides necessary power gain to produce adequate signal-to-noise ratio at the receiver. Up to 10 watts output from the power amplifier is available. A feedback loop provides a leveling voltage to the sweep oscillator so that transmitter power variations versus time or frequency can be eliminated. At the receiving site, a bandpass filter over the desired operating frequency range can be used, if necessary, to reject interfering signals. Consequently, at the output of the bandpass filter, undesired signals are reduced to the noise level. The power versus frequency at this point reflects the bandpass characteristics of the receiving antenna. A low-noise broadband amplifier is used before the detector to improve receiver sensitivity. The detector voltage versus time is integrated over one transmitter sweep period, and the integrated voltage represents average antenna power response over the transmitter sweep range. This integrated voltage is "read" at the end of the integration time and converted to logarithmic (dB) form. This converted value now represents average response of the reference antenna over the desired frequency range, expressed in dB form, at one spatial point. This value is temporarily stored until the above described process is repeated for the test antenna. After completion of this process for the test antenna, the two values are subtracted, and the result is the average gain of the test antenna expressed in dB which is referred to the average gain of the reference antenna over the same bandwidth.

In typical operation, the test antenna would rotate continuously while the transmitter sweep cycle is repeated each 10 msec. At an antenna rotation rate of 360 degrees per one-half hour, for example, the test antenna would move only 2×10^{-3} degrees during a transmitter sweep, and the antenna could be considered stationary. Thus, one obtains an essentially continuous spatial record of the average gain of the test antenna, referenced to the peak average gain of the reference antenna. If the absolute gain versus frequency characteristic of the reference antenna is known, its "average absolute" gain and the test antenna's "average absolute" gain can be calculated for any desired bandwidth up to the full sweep capability of the transmitter (typically one octave).

The following portions of this section present an analysis of the amplitude compensation approach, a system sensitivity analysis, hardware descriptions, and a discussion of applications of the Hybrid system.

B. AMPLITUDE COMPENSATION

If the power density of the signal which illuminates the test antenna varies with frequency and this received signal were integrated directly, an erroneous integrated gain would be obtained. Thus, either the incident power density must remain constant over the integration time or any variations must be compensated. As indicated previously, maintaining constant power density at the test antenna over an octave bandwidth would be difficult. Fortunately, electronic compensation can be achieved in a relatively straightforward manner.

The average gain of the test antenna can be determined from measurements of the average power of the antenna under test and of a reference antenna. Consider a general antenna test system consisting of a transmitter, a transmit antenna, a test antenna, and a receiver. The power received through the test antenna as a function of frequency may be expressed as

$$P_{test}(f) = kG_{TR}(f) \;\; P_T \;\; L(f) \;\; G_{test}(f) \quad , \qquad (2)$$

where

$P_{test}(f)$ = received power through test antenna as function of frequency,

k = constant,

$G_{TR}(f)$ = gain of transmit antenna as function of frequency,

P_T = transmitter power (assumed constant),

$L(f)$ = system losses (space attenuation) as function of frequency,

and

$G_{test}(f)$ = gain of test antenna as function of frequency.

It is desired to measure $P_{test}(f)$ and thereby determine the test antenna gain, $G_{test}(f)$. Therefore, all the other variables in the above equation must either be known as a function of frequency so that their effect can be removed mathematically, or the variables may be compensated electronically. Electronic compensation may be accomplished in real time, and it does not require knowledge of the functions which describe how each of the variables depend on frequency. Electronic compensation is accomplished through use of an additional reference antenna. The power received through the reference antenna, $P_{ref}(f)$, may be written as

$$P_{ref}(f) = k \; G_{TR}(f) \;\; P_T \;\; L(f) \;\; G'_{ref}(f) \quad , \qquad (3)$$

where $G'_{ref}(f)$ is the peak gain of the reference antenna as a function of frequency, and the other symbols are as previously defined. The hybrid system determines the average test antenna gain over some frequency range Δf. In terms of average quantities the two previous equations can be expressed as

$$\overline{P}_{ref} = k \; \overline{G}_{TR}\overline{G}'_{ref}\overline{L} \; P_T \text{ , and} \qquad (4)$$

$$\overline{P}_{test} = k\, \overline{G}_{TR} \overline{G}_{test}\, \overline{L}\, P_T \,, \tag{5}$$

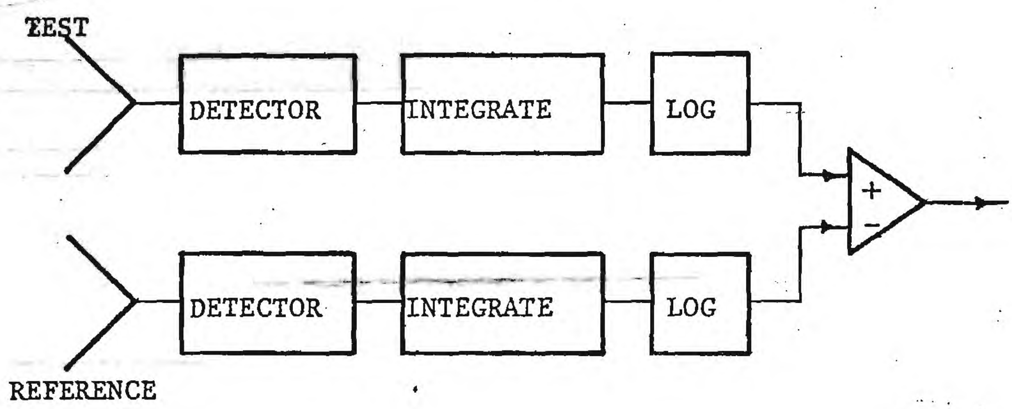
where the bar indicates an average over a specified Δf. Dividing $\overline{P}_{test}$ by $\overline{P}_{ref}$ and solving for $\overline{G}_{test}$ yields

$$\overline{G}_{test} = \overline{G}_{ref} \frac{\overline{P}_{test}}{\overline{P}_{ref}} \;. \tag{6}$$

Thus, average test antenna gain may be determined from measurements of $\overline{P}_{test}$ and $\overline{P}_{ref}$ if the average gain of the reference antenna is known. The effect of all other variables has been eliminated by taking the ratio, $\overline{P}_{test}/\overline{P}_{ref}$. The logarithmic or decibel form of the above Equation (6) is

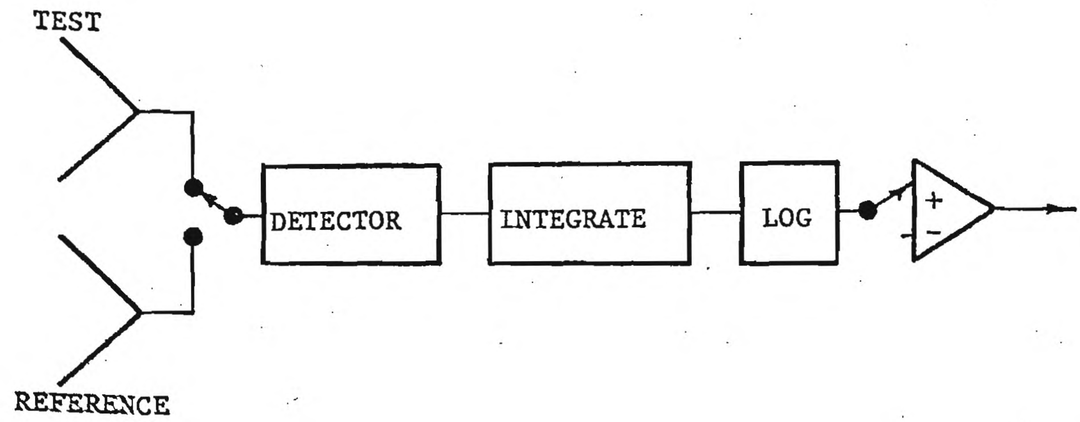
$$10 \log \frac{\overline{G}_{test}}{\overline{G}_{ref}} = 10 \log \overline{P}_{test} - 10 \log \overline{P}_{ref} \;. \tag{7}$$

The average gain of the test antenna, expressed in dB and referred to the average gain of the reference antenna, is obtained by taking the difference of the logarithms of the integrated received powers. The basic compensation concept* is shown in block diagram form as follows:



*An alternate compensation concept also was considered. The end result of this compensation concept was the variation of test antenna gain only, referred to the gain of the test antenna at one end of the frequency range. However, this concept required two electrically identical antennas and a voltage ramp at the receive site which was exactly time coincident over the entire sweep time with the transmitter frequency sweep ramp. In addition, this concept required constant transmitter power. Due to practical problems in meeting all these requirements, this concept is not attractive.

This circuit can be simplified and hardware savings can be realized by time-sharing one channel from the two antennas to the subtractor. The time-shared system becomes



On alternate transmitter sweeps, test and reference antennas are selected. It should be recognized that the log output must be "held" for one transmitter sweep. A completely analog implementation of the Hybrid system which incorporates this time-shared receiver channel has been designed, and it is discussed in following portions of this section.

C. SENSITIVITY ANALYSIS

1. Minimum Detectable Signal

The hybrid system is based on a readily realizable design which meets the measurement goals of a Concept I system, provides CW measurement capability, and will provide growth capability by incorporation of major portions of the transmitter into a Concept II system. Accordingly, meeting performance requirements with a relatively simple receiver design was a design goal which has been accomplished. As discussed in Appendix I, the most straightforward receiver design for obtaining broad instantaneous bandwidth (no tuning or LO tracking) with good sensitivity is the wideband RF receiver with a separate low-noise RF amplifier for each octave band.

The minimum detectable signal for a receiver is limited by thermal noise. Thermal noise power N is given by

$$N = k T_o B , \tag{8}$$

where

k = Boltzmann's constant,

T_o = reference temperature = 290^oK, and

B = receiver bandwidth.

The specific Hybrid system which is considered here is designed for the 2 to 4 GHz octave. Thermal noise in this 2-GHz band is given by

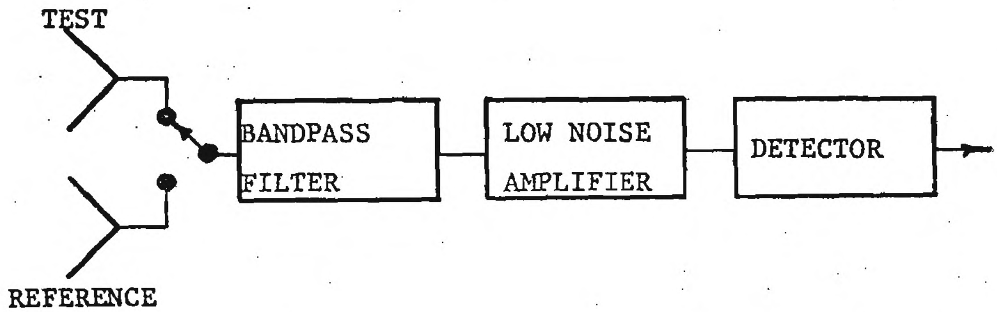
$$N = (4 \times 10^{-21} \text{ joules}) (2 \times 10^9/\text{sec})$$

$$= 8 \times 10^{-12} \text{ watts} = 8 \times 10^{-9} \text{ mW, or}$$

$$N = -81 \text{ dBm.}$$

Because typical semiconductor detectors have minimum detection levels on the order of -60 dBm, a signal level of -60 dBm would provide a 21-dB signal-to-noise ratio (SNR). However, since a SNR of 21 dB is not required, the detection of lower level signals can be achieved by use of a low-noise RF amplifier before the detector.

The effect of adding the RF amplifier, as illustrated in the receiver block diagram shown below, must be considered. For compensation, switching between the test and reference antenna is required. A bandpass filter has also been included for rejection of out-of-band interferring signals.



The switch and filter losses as well as the gain and noise figure of the RF amplifier must be considered in calculating sensitivity of this receiver configuration. These factors can be accounted for by the effective receiver noise figure, which is equivalent to the degradation in the signal-to-noise ratio (SNR) from the antenna terminals through the detector. This effective noise figure for this receiver configuration is

$$F_{eff} = L_s + L_f + (L_s + L_f)\,(F_{amp} - 1), \text{ where} \qquad (9)$$

L_s = switch insertion loss,

L_f = filter insertion loss, and

L_{amp} = noise figure of the RF amplifier.

The insertion losses of the filter and the switch are estimated at 1.0 dB each. For the 2-4 GHz range, typical amplifier specifications are the following.

Maximum noise figure:	4.5 dB
Minimum small signal gain:	20 to 35 dB
Power out @ 1-dB gain compression:	5 mW

With the above switch, filter, and amplifier specifications, the effective receiver noise figure would be

$$F_{eff} = 1.26 + 1.26 + (1.26 + 1.26)\ (1.82) = 7.65\ .$$

The noise figure, expressed in dB, is 8.84 dB.

Basedon the results obtained from Equations (8) and (9), the received signal power P_r at the antenna terminals can be determined if a minimum detection requirement of a 10-dB SNR is assumed. For an assumed SNR of 10 dB, the required signal power at the antenna terminals must be at least 10 dB above the effective noise level, $k\ T_o\ F_{eff}\ B$. Therefore, received power must be

$$\begin{aligned} P_r &\geq SNR\ (kT_o)\ F_{eff}\ B \qquad (10) \\ &= 10(4 \times 10^{-21})\ (7.65)\ (2 \times 10^{9}) \\ &= 6.12 \times 10^{-7}\ \text{mW, or} \approx -62\ \text{dBm.} \end{aligned}$$

With an RF amplifier gain of, say 25 dB, power into the detector would be a minimum of -37 dBm, or greater than 20 dBm above its noise level.

2. Transmitter Power Required

The transmitter design is based on providing up to 10 watts at the transmit antenna terminals so that adequate transmitter power is available for a variety of antenna conditions and for growth to the Concept II system. Antenna test ranges of up to 7000 feet are postulated. Besides the transmitter power and test range considerations, other factors which will affect the received power level are space loss, test frequency, and antenna size. Because the greater space loss occurs at higher frequencies, the received power for a 10-watt transmitter will be calculated for test operation at the high frequency edge of S-band, 4 GHz. Also, since a fairly directive transmit antenna should be used to minimize multipath problems, a 30-dB gain antenna will be assumed. For a typical aperture illumination function, the 3-dB beamwidth of this transmitting antenna will be approximately 5 degrees. A gain of 30 dB for the antenna under test will also be assumed. If smaller antennas were used, the test range could be considerably reduced. For example, the typical far-field range requirement ($R \geq 2\ D^2/\lambda$) for a 30-dB gain 4 GHz antenna is satisfied at a range of about 100 ft. Therefore, the above set of numbers represents an expected worst-case condition. The received power under these conditions would be

$$P_r = \frac{P_t\ G_t\ G_r\ \lambda^2}{(4\ \pi\ R)^2} \tag{11}$$

$$\approx 8 \times 10^{-2}\ \text{mW, or}$$

$$\left(P_r\right)_{dB} \approx -11\ \text{dBm},$$

where

P_t = power delivered to the transmitting antenna = 10 watts,

G_t = gain of transmitting antenna = 1000 (30 dB),

G_r = gain of receiving antenna under test = 1000 (30 dB),

λ = operating wavelength at 4.0 GHz = 0.25 feet, and

R = range between the two antennas = 7000 feet.

This -11 dBm is the received power in the main beam of the test antenna. Since -62 dBm of Equation (11) provides a 10-dB SNR, sidelobes of the test antenna as low as -51 dB below the peak of the main beam could be observed at a 10-dB SNR with the assumed transmitter power, test range, and antenna gains.

3. Detector Requirements

A minimum of 40-dB instantaneous dynamic range is required. This is approximately the maximum range which can be covered by a semiconductor detector, while maintaining square-law response, without special electronic compensation schemes. Typical semiconductor detectors cover the RF power range from 0 dBm to -40 dBm with square-law response and with corresponding detector output from 25 μV to 250 mV. The minimum received power at the test antenna terminals has been specified as -62 dBm. Switch and filter insertion losses have been estimated at a total of 2 dB maximum. Therefore, to obtain a minimum of -40 dBm at the detector input, the low noise amplifier must have a gain of at least 23 dB. Specifying a minimum gain of 25 dB would allow inserting a coupler for power monitoring and a limiter for receiver protection after the RF amplifier. Inserting these devices after the 25-dB gain low-noise amplifier will not materially degrade the effective noise figure of the receiver.

D. HARDWARE DESCRIPTIONS

Preliminary design of the hybrid system and component specification for the 2-4 GHz range have been completed. This subsection contains a description of the hardware necessary to implement this 2-4 GHz system.

1. Transmitter

A block diagram of the 2-4 GHz (S-band) transmitter is given in Figure 11. This transmitter, which will provide constant power at the 10-watt level across this 2 to 4 GHz range, consists of a basic signal source (solid-state sweep oscillator), power amplifier (helix traveling wave tube, TWT), and a power sampler to provide a leveling voltage for the sweep oscillator. Power output across the 2-4 GHz range is typically 10 mW $\pm$ 0.5 dB for solid-state sweepers. Thus, the TWT must have a gain of 30 dB to provide a 10-watt output. Octave bandwidth TWTs typically have small signal gain variations versus frequency. In a well constructed TWT, gross gain variations across the band can be held to about 3 dB with a fine grain variation about this gross curve of $\pm$ 1 dB. For the entire transmitter chain, power variations of 4-5 dB could occur. Thus, to achieve constant transmitter-power leveling is required.

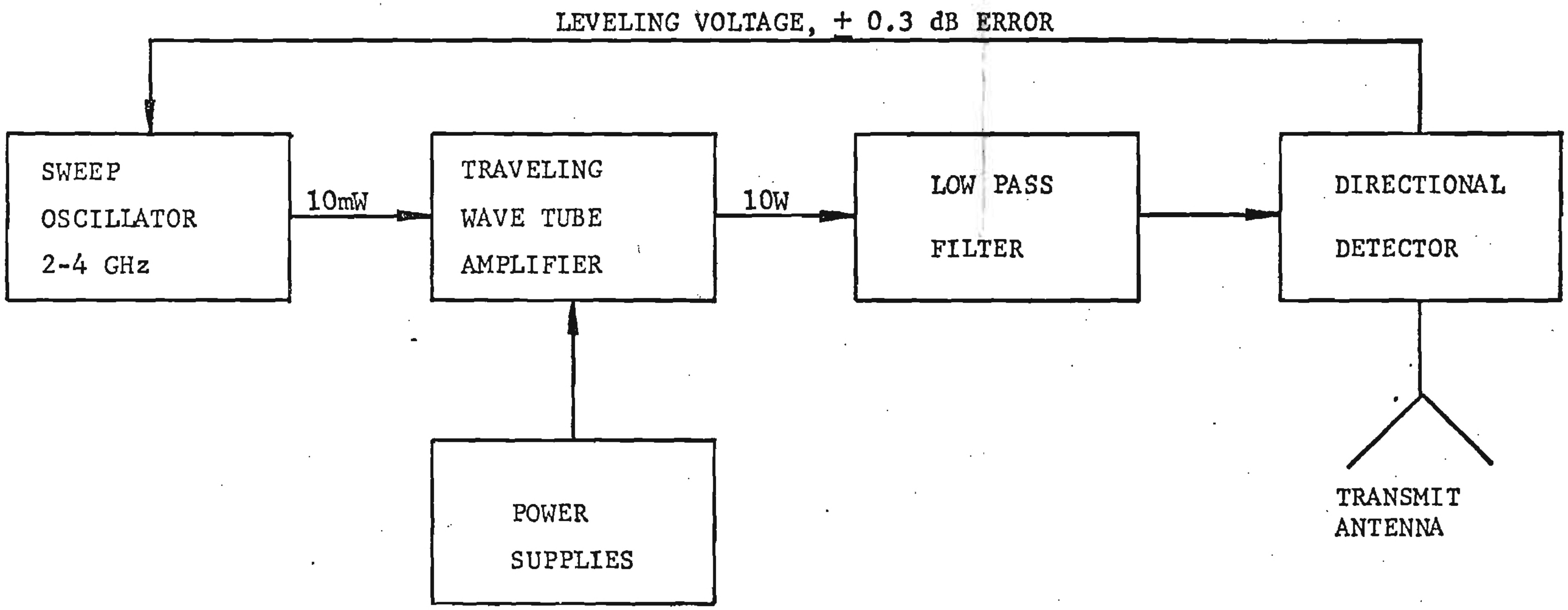


Figure 11. Hybrid system S-band transmitter block diagram

For long term reliability, solid-state sweep oscillators are preferred over conventional BWO (backward wave oscillator) sweep oscillators. For a typical 2-4 GHz sweeper, the output power variation with external leveling may be held to $\pm$ 0.1 dB plus leveling signal error. A TWT output of 10 watts will be obtained for an input of 10 mW if the TWT gain is 30 dB. If the upper limit of the TWT's small signal region is at 10 watts, more than a 10 mW output from the sweeper would drive the TWT into saturation. For signal purity, it is desirable to operate a TWT in the small signal region; therefore, a sweeper output level-set capability is desired. This level-set capability is also desirable to reduce the transmitter power to only that required for pattern measurements under varying range conditions. With a TWT noise figure of 35 dB and a gain of 30 dB, the noise power output in a 2 GHz band would be 56 dB below the 10-watt level so that a very broad dynamic range in the TWT output is available. (Noise power output may be calculated from Equation A-3 of Appendix I with the ENR term being equivalent here to the sum of gain plus noise figure.)

Remote programming is an option which many oscillators have available and is attractive for the hybrid system and for growth to a phase/amplitude system. Up to 1000 discrete frequencies may be remotely selected through a 12-line 8421 BCD input. A standard feature (RF Enable) permits the remote on/off control of the RF power. These features would allow remote selection of single frequencies for CW testing.

Octave bandwidth TWTs may have significant second harmonic output, depending on drive level and frequency. A low-pass filter is incorporated to remove any harmonics generated in the TWTA. Since low-pass filters are not very expensive ($40), the option of selecting a filter with the minimum possible cut-off for a particular sweep range is recommended. For example, if a 4 GHz low-pass filter were used when sweeping from 2 GHz to 3 GHz, the harmonic of the low end of the sweep range would not be cut-off. A 3-GHz cut-off would be appropriate, and a minimum of 2 low-pass filters per octave range is desirable. Pertinent specifications for the components of this transmitter are summarized in Table 2.

2. Receiver

The basic receiver design concept and performance of the critical components have been discussed in Subsection III-C, Sensitivity Analysis. A block diagram of the receiver is shown in Figure 12. This receiver accepts the RF signals from either the test or reference antenna and provides a dc output which is proportional to the power level of the RF signal. The receiver instantaneous dynamic range for square-law response is a minimum of 40 dB over the detector input range from 0 dBm to -40 dBm. The typical detector output voltage range for this power range is from 25 uV to 250 mV. However, this type of detector has sufficient sensitivity for use at power levels as low as -50 dBm, and as discussed later, the reference antenna compensation scheme tends to eliminate system non-linearities so that the useful dynamic range will be 50 dB or greater. Laboratory tests will be required to define the system accuracy over this detection range.

A limiter and power sampler (coupler) are incorporated for control of the maximum power level into the detector and for receiver calibration. Adjustment of the power level into the receiver is through the level-set attenuator in the sweep oscillator at the transmitting site. Characteristics of major components of the receiver are summarized in Table 3.

3. Data Processor

A block diagram of the data processor is given in Figure 13. The data processor is that part of the system which acts on the detector output

TABLE 2

COMPONENT SUMMARY* FOR S-BAND HYBRID SYSTEM TRANSMITTER

Sweep Oscillator	TWT Amplifier	Level Sensor
Nominal Power output: 10 mW	Small-Signal Output Capability: 10 watts CW	Frequency Range: 2-4 GHz
Frequency Range: 2-4 GHz	Nominal Gain: 30 dB	Output Voltage Flatness: $\pm$ 0.3 dB
Sweep Modes: Repetitive Sweep between Any Present Limits within 2-4 GHz; Fixed Bandwidth Sweeps about Selectable Center Frequency	Frequency Range: 2-4 GHz	Input Power: $\leq$ 10 W
Non-Sweep CW Mode	Beam Supply Regulation: $\pm$ 0.02%	Directivity: $\geq$ 20 dB
Remote On/Off Control	Helix Voltage Regulation: $\pm$ 0.1%	
Remote CW Frequency Selection	Line Voltage Variation: $\pm$ 10%	
Nominal Power Variation: $\pm$ 5 dB (Without External Leveling)	Interlocks/Interrupts: Time Delay; Beam/Helix Overcurrent; Thermal Cutout; Power Line Interrupt	
Sweep Linearity: $\pm$ 1%	Load VSWR: 2.5:1	
Sweep Time: Adjustable from 10 msec to 100 sec per sweep	Noise Figure: $\leq$ 35 dB	
Level-Set Attenuator		
External Level Accuracy: $\pm$ 0.1 dB plus level signal error		

*Nominal specifications are illustrated.

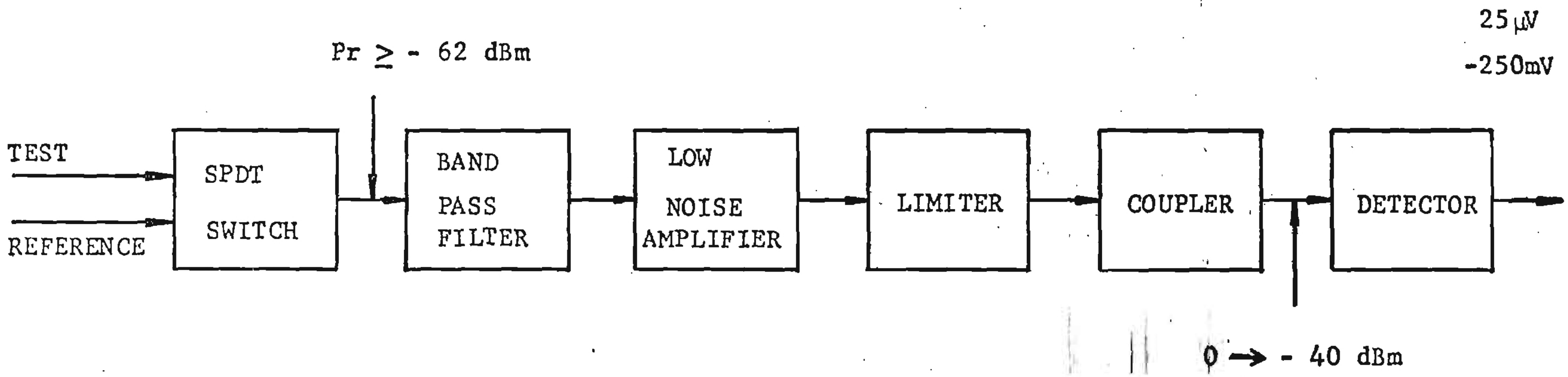


Figure 12. Hybrid system S-band receiver block diagram

TABLE 3

COMPONENT SUMMARY* FOR S-BAND HYBRID SYSTEM RECEIVER

Switch	Filter	Low-Noise Amplifier
Frequency: 2-4 GHz	Passband: 2-4 GHz	Type: Transistor
Maximum Insertion Loss: 1 dB	Rejection: 40 dB below 1.2 GHz and above 5.2 GHz	Frequency: 2-4 GHz
Switching Time: ≤ 0.5 μsec	Passband Maximum VSWR: 2:1	Maximum Noise Figure: 4.5 dB
Isolation: ≤ 40 dB	Insertion Loss: 1 dB max	Minimum Gain: 25 dB
Type: SPDT with integral driver		Gain Flatness: ± 1 dB
		Power out at 1 dB Gain Compression: ≥ 0 dBm
Limiter	**Coupler**	**Detector**
Frequency: 2-4 GHz	Frequency: 2-4 GHz	Frequency Range: 2-4 GHz
Power Handling: 1 W	Insertion Loss: 0.2 dB	Square Law Response: ± 0.5 dB from 0 dBm to -40 dBm
CW Leakage: ≤ 100 mW	Coupling: 20 dB	Minimum Useful Input· ≈ -50 dBm
Insertion Loss @ 0 dBm or less: ≤ 1.0 dB	Directivity: ≥ 20 dB	Voltage Sensitivity: 250 mV/mW
		Maximum Input: ≈ 100 mW

*Nominal specifications are illustrated.

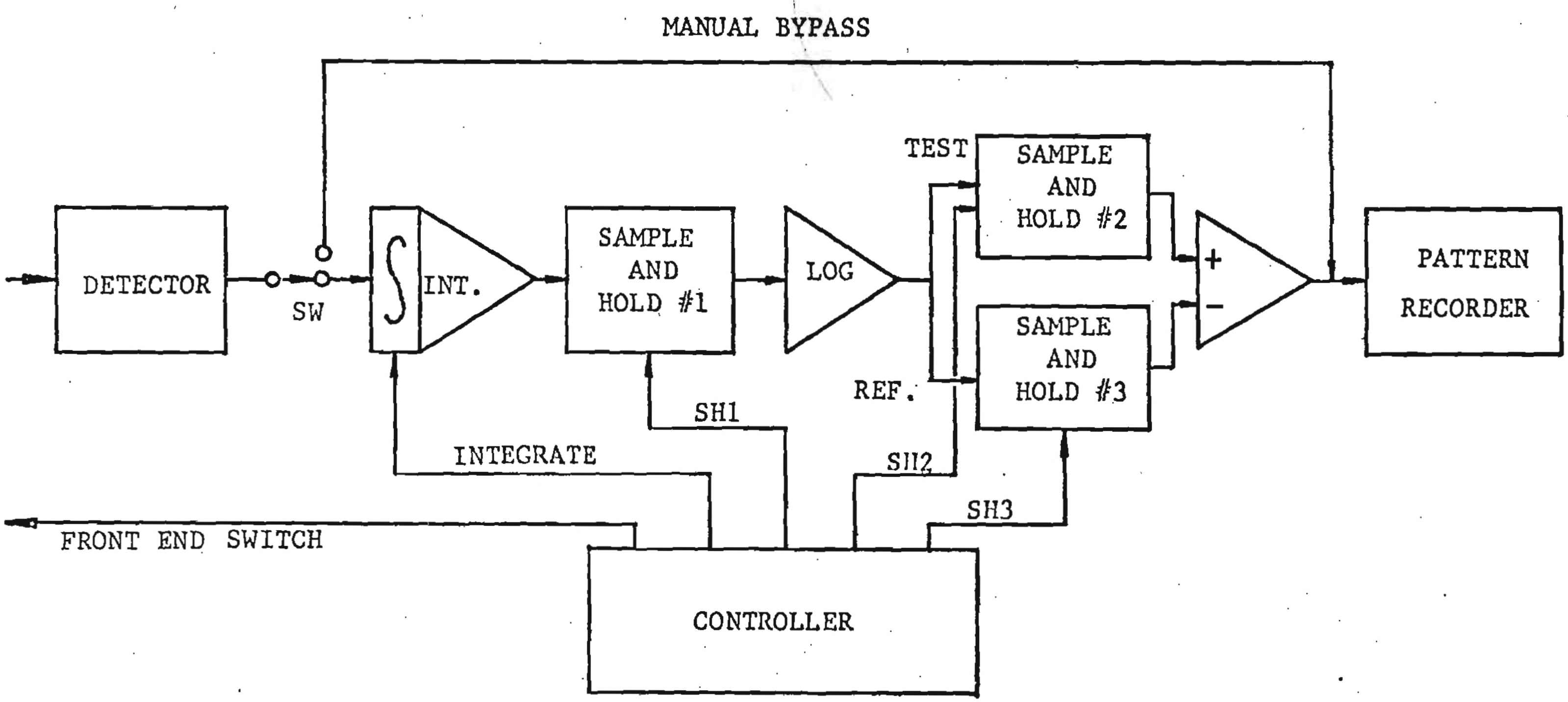


Figure 13. Hybrid system data processor block diagram

voltage to provide compensated average gain data. The time-varying signal from the detector represents the frequency dependent response of the antenna under test, plus variations due to any other frequency dependent elements in the entire measurement network from the transmitter through the detector. As previously discussed in Subsection B, the effect of all of these other variations is removed by amplitude compensation in the data processor. The data processor consists essentially of a circuit which integrates the detector output over one sweep period, a sample-and-hold device which samples the integrator output at the end of each sweep period and holds this value until the logarithm can be performed by the log amplifier. The logarithm signal is fed to a sample-and-hold network combined with a subtractor. One of the sample-and-hold modules in this network always contains the most recent value of the logarithm of the reference antenna's average response over the desired frequency range, while the other sample-and-hold contains the most recent value of the logarithm of the test antenna's average response over this sweep range. The subtractor output is the difference of these two log signals stored in the sample-and-hold circuits. This difference (which corresponds to a power ratio) is the integrated (average) response of the test antenna, which is expressed in dB and is referred to the average response of the reference antenna.

A typical transmitter sweep time for the maximum data recording rate would be 10 msec. Therefore, sample-and-hold (S/H) Module #1, which contains the integrator output, must be updated each 10 msec. Since the test and the reference antennas are selected on alternate transmitter sweeps, sample-and-hold (S/H) Modules #2 and #3, which contain the test and reference antenna responses, respectively, must be updated each 20 msec. Since the reference antenna will be stationary, the output of S/H Module #3 will remain essentially constant for a given frequency sweep range.

Very large antenna positioners have maximum scan rates on the order of 0.5 rpm. Even at this maximum scan rate, in 20 msec the test antenna would rotate through only 0.060 degree. Broadband average antenna response will not change dramatically over this angular increment (broadband pattern characteristics are discussed in the following section). Therefore, even for the maximum angular rotational speed of the antenna positioner, there would not be appreciable changes in successive values of either S/H Module #2 or S/H Module #3 output, and the subtractor output would be a smooth function of time (or test antenna spatial coordinate). The impact of various test antenna characteristics on spatial rotation rate is also discussed in the following section.

The integrator is shown schematically in Figure 14. The integrator consists of a commercially-available low-noise (3 μV rms) operational amplifier plus the appropriate external resistor and capacitor to obtain the required response time. The output voltage e_o is given by

$$e_o = \frac{1}{RC} \int_o^{t_1} e_i(t)dt \quad , \tag{12}$$

where R and C are defined by Figure 14, t_1 is the integration time, and $e_i(t)$ is the input voltage to the integrator. Consider a typical situation diagramed in Figure 15. The upper sketch shows a waveform which might appear at the output of the detector. A voltage which varies in time between $\approx$ 0 V and 250 mv is shown. The area under the curve represents an average gain-bandwidth measure of the antenna under test so that the integral of this voltage, which represents the gain-bandwidth measure of the test antenna, might appear as shown in the lower sketch of Figure 15.

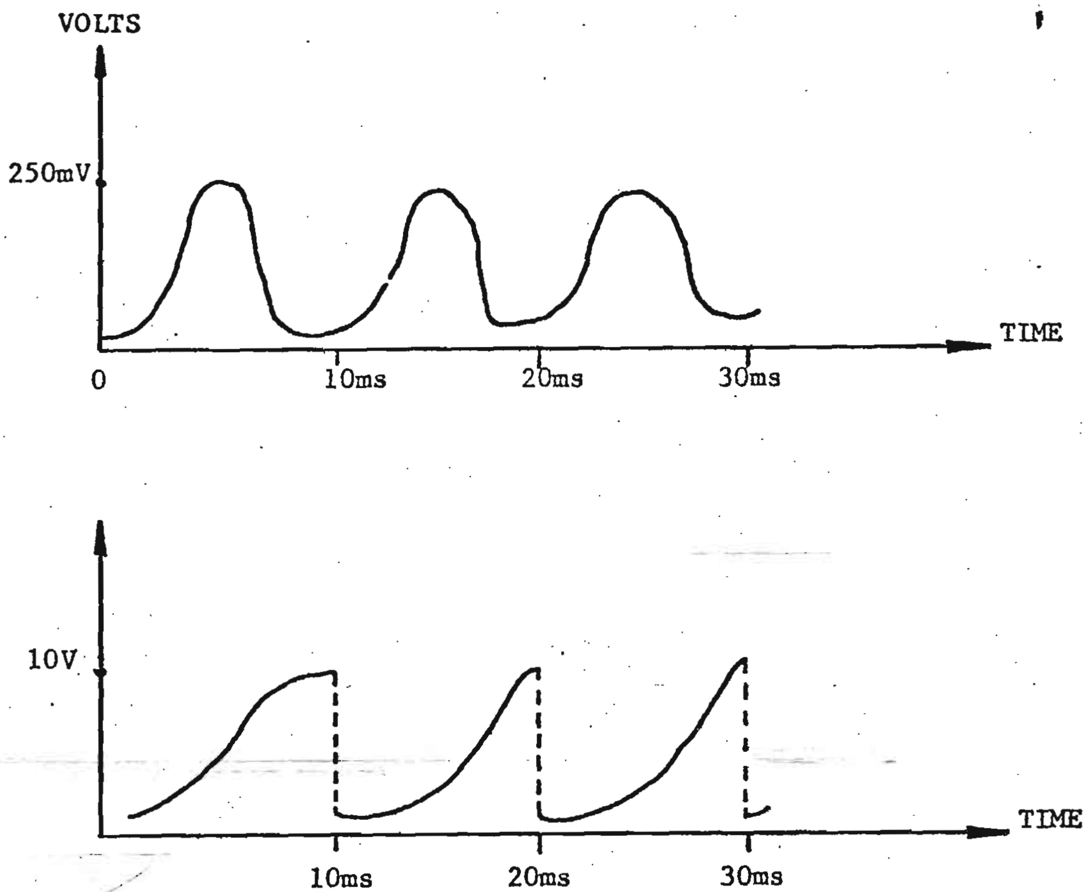


Figure 15. Typical detector and integrator waveforms

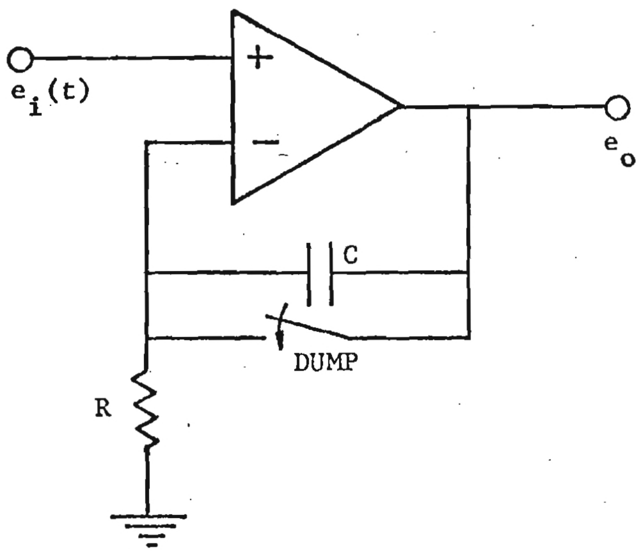


Figure 14. Integrator for Hybrid system data processor

If $e_i(t)$ has a cosine-type variation over the 10 msec measurement interval and a peak value of 0.250 volt, then $e_i(t)$ can be represented by

$$e_i(t) = 0.125\ (1 - \cos 200\pi t). \tag{13}$$

The value at the output of the integrator after a 10 msec period is given by

$$e_o = (1/RC) \int_0^{0.01} 0.125\ (1 - \cos 200\pi t)\ dt = 0.00125/RC\ . \tag{14}$$

It is desirable that the integrator output voltage e_o be in the optimum range of 1 mv to 10 volts, which is necessary for the subsequent operation of the S/H Module #1 and log amplifier. Therefore, the RC product must be selected and controlled by the operator in order to provide an optimum voltage range for a wide range of frequency sensitive antennas that may be tested. For the present example involving a cosine-type detector output function, an integrator output of 10 volts is desired when the input voltage covers the 0 - 250 mv range. From Equation (14), the RC factor is calculated to be

$$RC = .00125/10v = 1.25 \times 10^{-4}\ .$$

Therefore, selecting $R = 10^4$ ohms and $C = 1.25 \times 10^{-8}$ farads will satisfy the condition on the output voltage range.

The sample-and-hold Module #1 must sample the integrator output at the end of each integration period and hold that value until the end of the following integration period. Standard commercially available modules that are well suited for this type of application are available. For example, one suitable unit has an acquisition time (time necessary to acquire the input signal and approximately settle to its steady-state value) of 15 μsec and an output voltage droop factor of 5 μV/msec. This value of acquisition time is short compared to changes in the integrator output so that no error is introduced. Although the droop factor must be considered, a voltage droop of 5 μV/msec over a 10 msec period will only be 50 μV. Therefore, because the signal voltages cover four decades (40 dB) from 1 mV to 10 V, the maximum percentage droop will occur for the minimum signal (1 mV) and it will be only 5%.

The accuracy and dynamic range of standard logarithmic amplifiers (log amp) can be considerably improved by operating them in a current input mode rather than in a voltage input mode. Consequently, a voltage-to-current converter (VCC) is utilized as the input element to the logarithmic amplifier, as shown in Figure 16. Both the operational amplifier in the VCC and the entire log amp are standard commercial products. The voltage to current conversion is expressed by

$$i_L = -e_i/R_2, \text{ provided} \tag{15}$$

$$R_3/R_2 = R_F/R_1 ,$$

where

i_L is the current input to the log amp,

e_i is the voltage input to the VCC, and

R_2, R_3, R_F, and R_1 are indicated in Figure 16.

Four decades of input current (corresponding to the four decades of signal voltages of 1 mV to 10 V) to the log amp are required. Since the midrange of a typical log amp's input capability is from -10^{-4} amperes to -10^{-8} amperes, and since e_i will range from 1 mV to 10 V, from Equation (15) R_2 should be 10^5 ohms. The log amp output voltage range will be from -6 V to +2 V for the 10^{-8} amperes to 10^{-4} amperes current input, respectively. This -6 V to +2 V is shifted into the 0-10 V range by an operational amplifier with a fixed bias input.

Sample-and-hold Modules #2 and #3 are identical to S/H Module #1. At the end of the test and reference antenna integration intervals, S/H Module #2 and S/H Module #3, respectively, sample the log output and hold the values

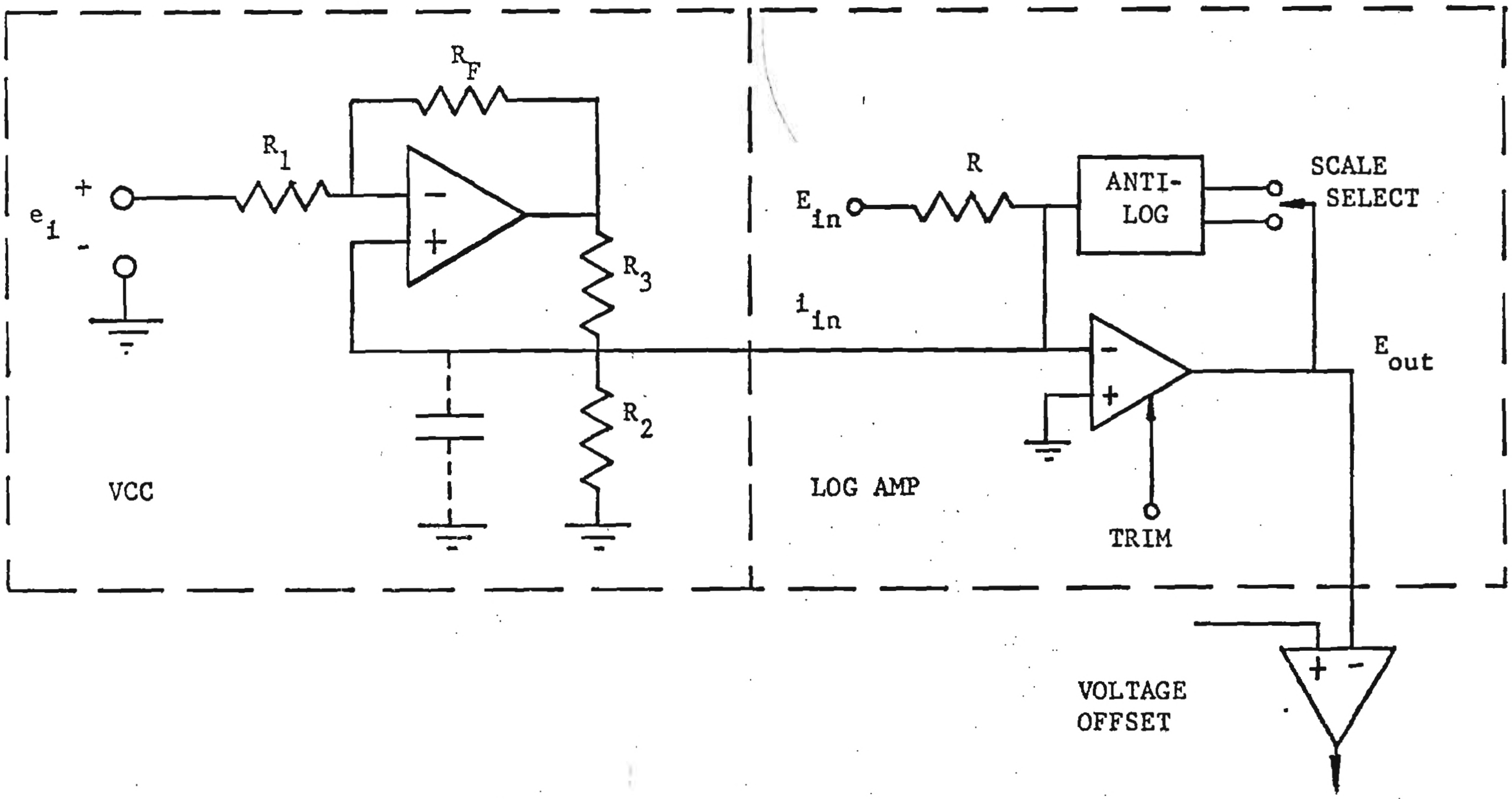


Figure 16. Logarithmic amplifier with voltage-to-current converter

until each respective output is updated on alternate transmitter sweeps. The output value of the subtractor, which is a standard operational amplifier module, typically ranges from 0 to 10 volts.

For use with the Hybrid system, the Scientific-Atlanta 1520 Rectangular Pattern Recorder would be equipped with the Series 1556 dc-chopper preamplifier. When so equipped, dc input voltages of up to 100 volts are permitted. Full scale sensitivity is 0.01 volt and the noise level is 10 μV rms. Thus, the 0-10 V input signal will interface well with this recorder.

Characteristics of major components of the data processor are summarized in Table 4. It should be noted that the data processor is not frequency dependent, and it can be used for various RF ranges.

4. Timing and Control

Five control lines are required to synchronize and control the operation of the various data processor modules. These lines control the front end switch, the integrator, and the three sample/hold modules. The front end switch selects either the calibrated reference antenna or the test antenna. The integrate signal starts and stops the integration of the detected video signal, and the sample/hold lines select and store the appropriate signals from the integrator and the log amp.

In Figure 17, each of the signals between processing modules is illustrated schematically in the top half of the figure, and the control signals on each of the five control lines are shown in the bottom half of the figure. A transmitter sweep rate of 100 Hz is indicated by the video signal directly from the detector. The reference antenna integrated response is stored at t = 10 ms, and the test antenna integrated response is stored at t = 20 ms. The control signals SH2 and SH3 transfer the voltages at the log amp output to either the subtractor (+) port or the subtractor (-) port. These inputs

TABLE 4

COMPONENT SUMMARY* OF HYBRID** SYSTEM DATA PROCESSOR

Operational Amplifier (1)	Sample-and-Hold	Logarithmic Amplifier
Output: $\pm$ 10 V @ 5 mA	Acquisition Time: 5 μsec to 0.01% of steady state	Rated Output: $\pm$ 10 V @ 5 mA
Input Impedance: 10^{11} ohms (3.5 pF)	Droop Rate: 5 μV/msec	Accuracy of Log: $\leq$ 1%, referred to input
Input Noise: Voltage, 3 μV rms; Current, 0.1 pA peak-to-peak	Input Range: $\pm$ 10 V	Dynamic Range: 60 dB, 10^{-9}-10^{-3} amp
Input Voltage: +8 to -10 V	Input Impedance: 10^{9} ohms	
Operating Temperature Range: 0 to +70°C	Aperture Delay: 40 nsec	
	Aperture Jitter: 4 nsec	

*Nominal Specifications are illustrated.
**Used in integrate, log, and subtract functional modules.

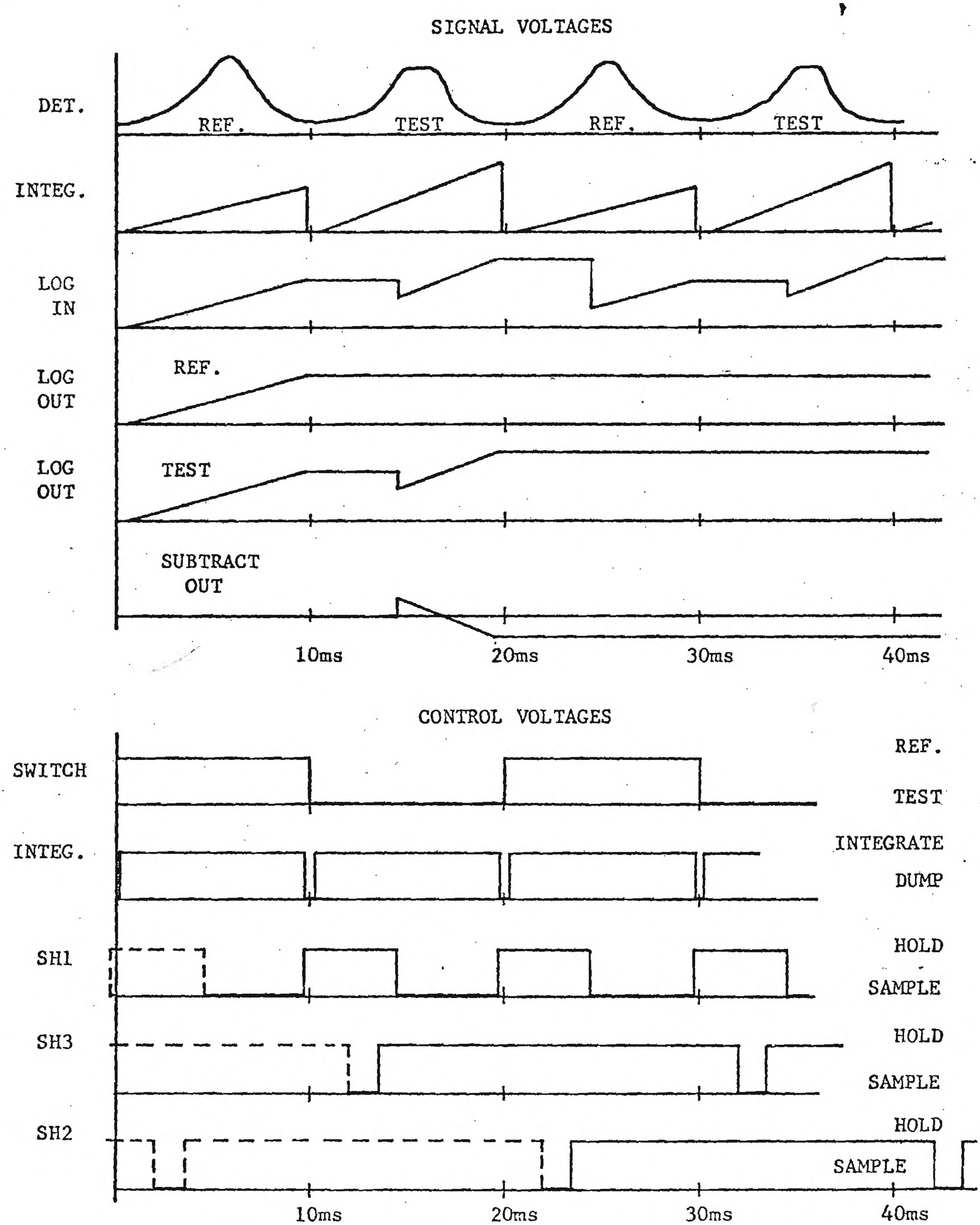


Figure 17. Illustration of Hybrid system timing and control requirements

at the subtractor lead to the self-calibration of the measured data on a sweep-by-sweep basis.

The five control signals are derived from the basic square wave signal controlling the front end switch. The integrator control signal, SH2, and SH3 are simple monostable multivibrator output signals which are triggered with appropriate delays relative to the front end switch signal. SH1 is a square wave at twice the rate of the square wave which controls the front end switch.

The front-end-switch signal may be generated in several ways. The simplest way, conceptually, is to derive the switch signal from the transmitter sweep waveform by transmitting the timing signal via a direct data link of suitable capacity or by telephone service with modems at both sites. However, for mobility and field use it is desirable that operation of the Hybrid system not be dependent on such a link, and an alternate approach has been selected. With the alternate approach, the basic timing signal is synchronized with the test and reference transmitter sweeps by detecting the beginning of the sweep cycles from a sample of the receiver signal. This detected sample drives a Schmidt Trigger circuit which forms a rectangular pulse which is shaped to be a square wave and is appropriately delayed to drive the front end switch between transmitter re-sets. Synchronization is maintained by detecting the leading edge of each transmitter sweep cycle.

Pushbutton switches for manual selection of the test or reference antenna position are required for the single frequency mode of operation.

5. Transmit and Reference Antennas

As described in Appendix I-C, broadband multipolarized antennas suitable for the hybrid system transmit and reference antennas are available. For a feasibility demonstration, it is expected that the initial 2-4 GHz system will employ relatively simple transmit and reference antennas and that they will be "identical." A broadband linearly polarized feed with a paraboloid reflector is a practical choice which can meet the requirements of this demonstration system. Furthermore, the system could be upgraded to a multipolarized system by replacement of the feeds.

As indicated in the sensitivity analysis, the reflector should be sized to provide a gain of approximately 30 dB at 4 GHz. The power gain of a circular aperture with a typical illumination function (-25 dB first sidelobe) can be expressed as

$$G = K \frac{A \, 4\pi}{\lambda^2} \tag{16}$$

where

A = aperture area,

λ = wavelength, and

K = efficiency factor.

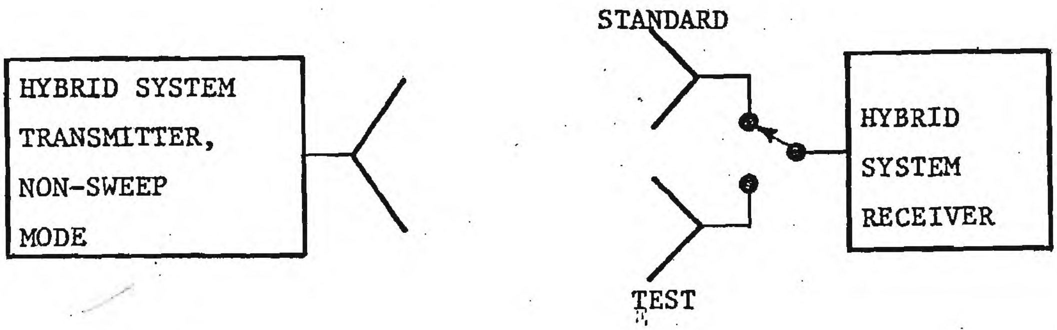
Expressing aperture area in terms of diameter d, and solving for d at a wavelength of 2.95 inches (f = 4 GHz), an efficiency of 0.5, and a gain of 1000 (30 dB), one gets $d \approx 42$ inches. The 3-dB beamwidth achieved with this illumination function is about 5 degrees at 4 GHz.

The antenna feed can be simply an open ended ridged waveguide with a waveguide-to-coax adapter for interface to the transmitter and receiver. The use of ridged waveguide will offer improved broadband matching over that obtainable with conventional rectangular guide. For example, a VSWR of 1.06:1 is typical for a coaxial to dual-ridged transformation over the 2-4 GHz range.

This simple antenna system will provide a cost-effective approach for realization of the initial 2-4 GHz hybrid system.

E. CALIBRATION

Since the Hybrid system provides integrated broadband test antenna data which are referred to the broadband response of a reference antenna, proper interpretation of the test data requires knowledge of the broadband properties of the reference antenna. The reference antenna can be calibrated versus a set of standard gain antennas covering the measurement range of the particular Hybrid system. Calibration of a 2-4 GHz reference antenna is described here. Calibration for other frequency ranges would be similar. The experimental setup is sketched below.



The reference antenna should be calibrated (have its absolute gain measured) at small frequency increments across the frequency range for which it is to serve as the reference. Since the gain of the reference antenna may not be a linear function of frequency, it is important to have a relatively large number of samples from which to determine the average gain over any desired bandwidth. Calibration of the reference antenna gain at each 100 MHz interval across the 2-4 GHz range will be assumed. Since standard gain horn antennas cover waveguide bands, two standards will be required to cover the 2-4 GHz range: one covering from 2.0 to 2.60 GHz, and one covering from 2.60 to 4.0 GHz. Each standard would be supplied with a gain curve showing its gain across its entire bandwidth.

The reference antenna would then be calibrated as follows. The antenna which is to be calibrated as a reference and one of the standard gain horns are mounted at the receiver site. The Hybrid system transmitter and receiver are operated in their non-sweep mode, sequentially covering a set of frequencies separated by 100 MHz across the range of the standard gain horn. At each of these frequencies, the gain of the antenna under calibration is determined relative to the gain of the standard. This procedure is followed with the second standard also until a set of absolute gain values is obtained for the reference antenna every 100 MHz ($G_{2.0}$, $G_{2.1}$ . . . $G_{3.9}$, $G_{4.0}$) across the 2-4 GHz range. Now, when this calibrated antenna is used as the reference in broadband measurements on an unknown antenna, its known gain at some frequency within the desired measurement range is used as the reference to convert measured data to absolute data. Average gain could also be calculated by converting the dB data into linear form and taking their average. The average gain of the reference antenna across the 3.0 to 3.5 GHz range would be

$$\bar{g}_{ref} = \frac{g_{3.0} + g_{3.1} + g_{3.2} + g_{3.4} + g_{3.5}}{5} , \tag{17}$$

where g represents a respective gain in absolute (non-dB) units.

F. SYSTEM ACCURACY

Since the logarithmic amplifier is expected to be the least accurate element in the hybrid system, the impact of errors in the log module has been examined. As indicated previously, a consequence of the electronic compensation scheme is the improvement of overall system accuracy. This improvement is illustrated by the following analyses of the impact of log module errors on system accuracy.

The log module ideally performs the following transform

$$V_{out} = -2 \log_{10} \left(V_{in}/V_{ref}\right) , \tag{18}$$

where V_{ref} is a fixed reference voltage supplied to the log amp and V_{in} is the input voltage. The transform error is taken into account by the following equation.

$$V_{out} + \epsilon_o = -2 \log_{10} \left[(V_{in} + \epsilon_i)/V_{ref} \right] , \tag{19}$$

where ϵ_o accounts for departures from ideal V_{out} for various errors ϵ_i in the input. The error at the output of the log module (effectively, the input to the subtractor) has been calculated for input errors of $\pm$ 5% and $\pm$ 10% and tabulated in Table 5 for a reference voltage of 100 mV.

In this error analysis, log amp errors of $\pm$ 10% and $\pm$ 5% referred to the log input voltage were assumed. The system output error was calculated at representative points across the 1 mV to 10 V log module input range. The assumptions for this estimate are the following:

(1) the log module error is defined in terms of the input voltage to the module,

(2) the overall processor error is defined in terms of the output voltage from the subtractor,

(3) no errors are considered except log module error,

(4) for worst case conditions, a positive extreme error is assumed for 1 mV input, no error is assumed for 100 mV input, and negative extreme error is assumed for 10 V input (40 dB dynamic range of interest), and

(5) the log module transfer function is stationary, not time-varying.

The error at the output of the log module (effectively, the input to the subtractor) has been calculated for input errors ϵ_i of $\pm$ 5% and $\pm$ 10% and tabulated in Table 5 for a reference voltage of 100 mV.

The subtractor output voltage is equal to the difference between successive measurements of the gain-bandwidth product for the test and reference

TABLE 5

ACTUAL LOG AMP OUTPUT VOLTAGE

V_{in}	$V_{out} + \epsilon_o$				
	$\epsilon_i = 0$	$\epsilon_i = +5\%$	$\epsilon_i = -5\%$	$\epsilon_i = +10\%$	$\epsilon_i = -10\%$
1 mv	+4	3.958	4.045	3.917	4.092
100 mv	0	- .042	.045	- .083	.092
10 v	-4	-4.042	-3.955	-4.083	-3.980

TABLE 6

SUBTRACTOR OUTPUT VOLTAGES

V_2	ϵ_i	$V_1 \rightarrow$ 1 mV ($\epsilon_i = +10\%$)	100 mV ($\epsilon_i = 0$)	10 V ($\epsilon_i = 10\%$)
1 mV	+10%	0	-3.917(2.1%)	-7.825(2.2%)
100 mV	0	3.917(2.1%)	0	-3.908(2.3%)
10 V	-10%	7.825(2.2%)	3.908(2.3%)	0

antennas. The ideal equation for the subtractor output is

$$S_{out} = -2 \log_{10} (V_1/V_{ref}) + 2 \log_{10} (V_2/V_{ref}), \tag{20}$$

where V_1 and V_2 represent the integrated detector voltages due to the reference antenna and the antenna under test, respectively. Based on the worst case assumption (4), the subtractor output voltage and the percent error for each error case have been calculated. These data are given in Table 6.

The errors are expressed as a percent of subtractor output with no errors in the log module. It may be seen that with the same gain-bandwidth product for the test and reference antennas ($V_1 = V_2$), there is no error in subtractor output, regardless of the percent of log transform error. The maximum error (2.3%) occurs for the maximum difference in gain-bandwidth response of the two antennas. The log transform error of 10% is larger than that expected when operating the log module in the current mode under nominal conditions. However, changes in operating temperature, bias voltage drifts, etc., could lead to errors of this magnitude. Even under these conditions, the error in the system output is not large. Laboratory tests will be required to establish actual system accuracy under a variety of conditions. Under all conditions, when V_1 and V_2 are equal, no error occurs in the subtractor output.

To obtain maximum accuracy from the Hybrid system, there are certain operational constraints which must be observed. The system has been designed for optimum operation over specific power levels and dynamic ranges. For example, the power level into the detector should not be more than 0 dBm maximum, at the peak gain of the test and reference antennas. In order to avoid saturating the system, the transmitter power level should be reduced to meet this requirement. In addition, the peak detector output voltages should be approximately equal for the test and the reference antennas. This requirement means that the test and reference antennas should be oriented for their maximum gains, and the power level equalized through use of a broadband attenuator immediately following the higher gain antenna. Of course, this attenuation value must be accounted for in converting the measured relative gain into absolute gain.

As indicated previously, the Hybrid system is configured to accommodate a typical maximum positioner rotation rate of 0.5 rpm with a resulting spatial uncertainty of 0.030 degree in a 10-msec sweep time. If more precise spatial data are desired, the antenna rotation rate can be reduced with an increase in data-recording time. For example, if the rotation rate is reduced to 360 degrees/one-half hour, the resulting spatial uncertainty is 0.002 degree.

G. DATA

The Hybrid system is designed to produce broadband average gain and pattern data as well as CW data at any selected frequency within its operating range. For single frequency operation, the sweep oscillator is switched to CW and tuned to the desired frequency. (Tuning may be either manual or remote if a data link between the transmit and receiving sites is available.) At the receiver site, timing of the processor functions continues but transmitter sync is not required. The basic 100-MHz timing signal is generated by a free running oscillator. This timing feature and CW operation are selected by pushbutton control and data to the pattern recorder remains in dB form and is fully compensated and normalized to the reference antenna.

In its primary mode of operation, the Hybrid system produces average gain/pattern information over a selected bandwidth. This bandwidth can be from zero up to an octave. The type of data produced is discussed and illustrated in Section IV-A, which is concerned with applications of swept-frequency phase/amplitude systems.

SECTION IV

CONCEPT II SYSTEMS

A. INTRODUCTION

Concept II systems are based on the use of a broadband tunable signal source and a synchronously tuned receiver to provide sampled data which can be either displayed in real time or stored on magnetic tape for diagnostic processing. Because of the large amounts of data which are involved, Concept II systems lend themselves naturally toward an automatic computer-controlled implementation. Initially, both amplitude-only and amplitude-plus-phase Concept II systems were considered. However, it was found that except for no receiver requirement to measure phase, the implementation of a Concept II swept-frequency sampled-data amplitude-only system requires equipment similar to that needed for the implementation of a Concept II phase-plus-amplitude system. However, because broadband amplitude-only data are provided by the less complex Hybrid system, development of a Concept II type of system would not be an effective utilization of resources for instrumentation if the end item is to be used strictly for obtaining amplitude-only data. Therefore, only the phase-plus-amplitude Concept II systems were further investigated.

Phase information can be important in determining the performance of very broadband systems. However, the importance of the phase information depends on the application. For example, both phase and amplitude information would be necessary in determining the antenna effect upon pulse shape or in defining pulse compression performance limitations. A new area in which phase information would be useful is the application of measured target backscatter phase characteristics to aid in target identification. In addition, if the antennas phase properties are known, compensation networks can be added to achieve more nearly ideal system performance. Areas where phase information is not required is in determination of broadband average gain and broadband spatial patterns. An application of the broadband spatial pattern is to determine the received power from a sidelobe noise jammer so that the system signal-to-noise (interfering) ratio can be found. For this application, the RMS sidelobe levels over the frequency band of the interfering signal may be used to compute the interference power level. The RMS sidelobe levels can be determined over the desired bandwidth from amplitude-only measurements. Complete amplitude and phase information in the sidelobe regions is required if the antenna effect on the pulses received through a sidelobe is of interest. Since the distortion of pulses received through the sidelobes is not usually of interest, because the radar system is designed for main-beam operation, it appears that sidelobe phase information is of limited practical value.

The types of antenna analyses which can be performed using amplitude-only and amplitude-plus-phase data can be examined analytically. If amplitude-only data are used, mathematically the patterns which are obtained are equivalent to those which would be obtained with a band limited white noise measurement system. The spatial pattern for this case may be expressed as

$$P(\theta, \Delta\omega) = \int_{\Delta\omega} |G_v(\omega,\theta)|^2 \, \Phi(\omega) \, d\omega , \tag{21}$$

where

$G_v(\omega, \theta)$ = complex voltage gain, and

$\Phi(\omega)$ = incident signal spectrum.

That is, the average received power over a band of frequencies can be obtained by integrating the power gain of the antenna over the frequency band of interest, weighted by the incident-signal spectrum. Of course, the antenna gain at a single frequency, or as a function of frequency for any given spatial angle, can also be extracted from measured amplitude data. Figures 18 through 20 show the effect on the antenna spatial pattern of increasing the integration bandwidth for a rectangular aperture antenna with uniform illumination. For small bandwidths, the pattern does not deviate significantly from the monochromatic pattern. However, as the bandwidth is increased, the pattern nulls are obscured, and the sidelobe peaks approach their average values. This type of information from amplitude-only data can be used in determining the susceptibility of the antenna to sidelobe noise jammers.

In order to account for pulse distortion effects, the complex voltage gain of the antenna must be known as a function of frequency. From these data, the received pulse waveform can be computed if the spectrum of the incident pulse is known. Thus, pulse distortion affects can be computed as a function of spatial angle. This may be expressed as

$$V_{rcv}(\theta, t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} G_v(\theta, \omega) \ \Phi_i(\omega) \ e^{j\omega} \, d\omega \tag{22}$$

where

V_{rcv} = received pulse voltage at spatial angle θ, as a function of time,

$G_v(\theta, \omega)$ = complex voltage gain of antenna,

$\Phi_i(\omega) = \int_{\infty}^{\infty} \varphi(t) \ e^{-j\omega \ t} dt$, and

$\varphi(t)$ = incident pulse.

The antenna effect on various pulse shapes φ (t) can be computed from the above equation after the required amplitude and phase data are measured for

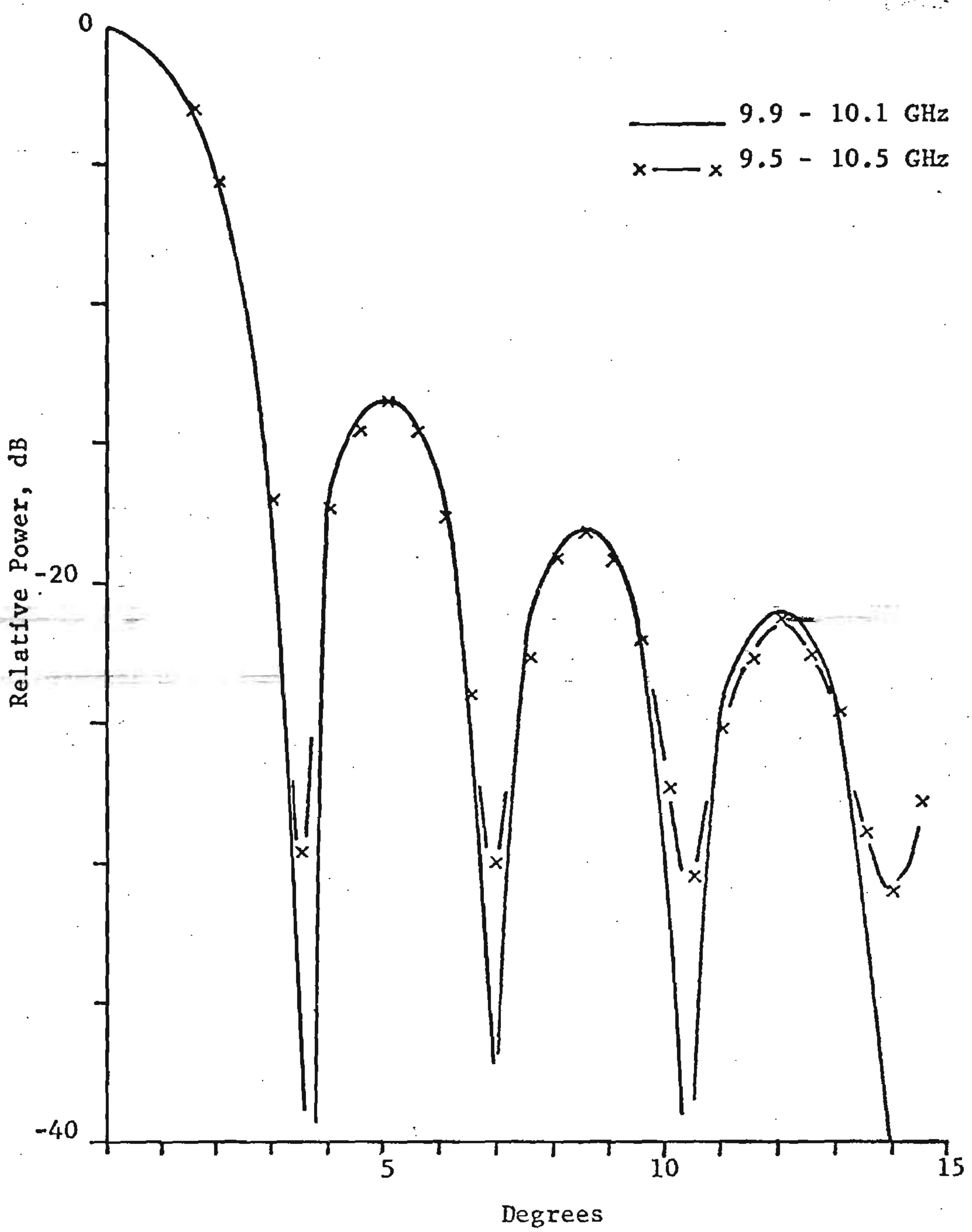


Figure 18. Broadband pattern for uniformly illuminated rectangular aperture, 1 GHz bandwidth

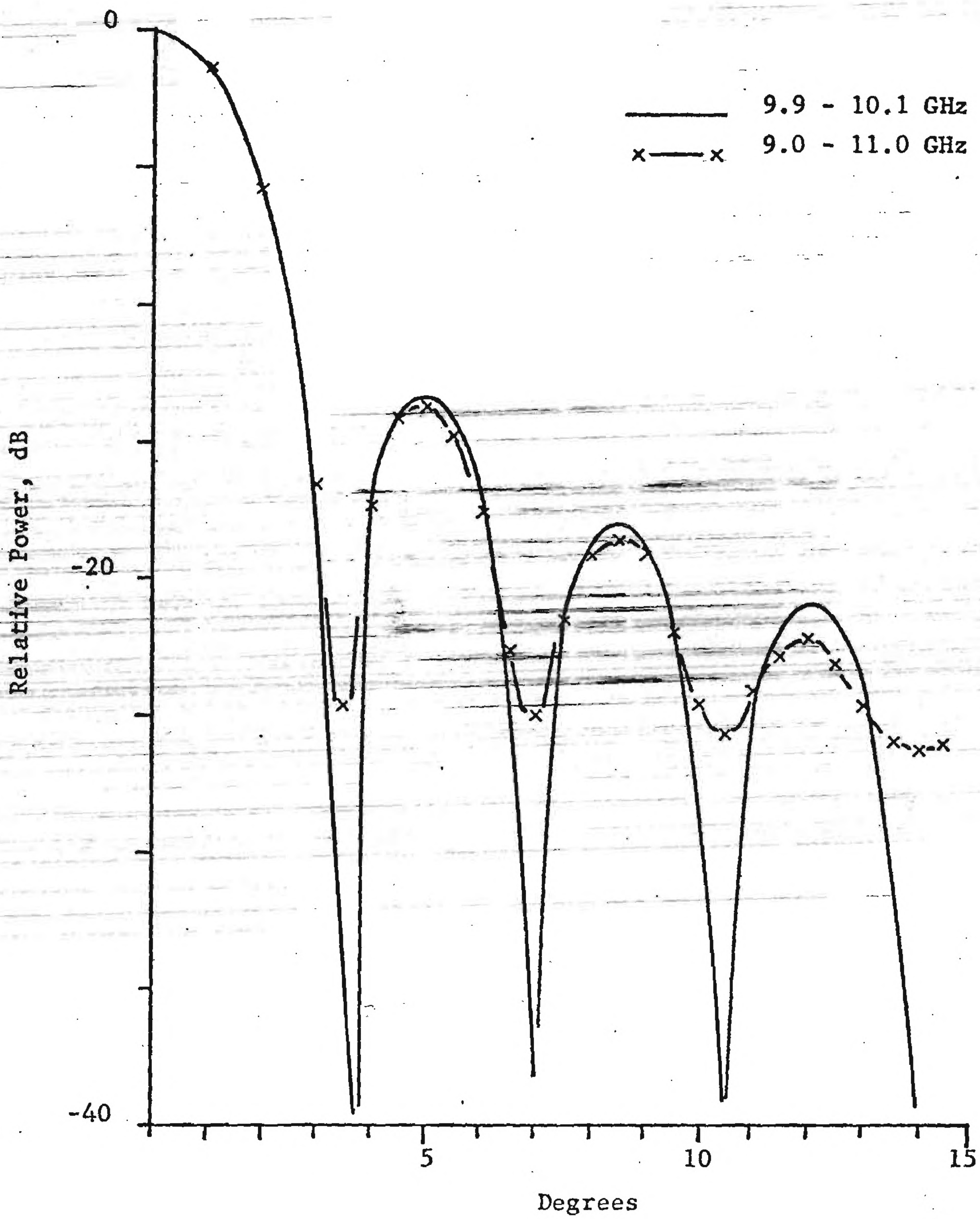


Figure 19. Broadband pattern for uniformly illuminated rectangular aperture, 2 GHz bandwidth

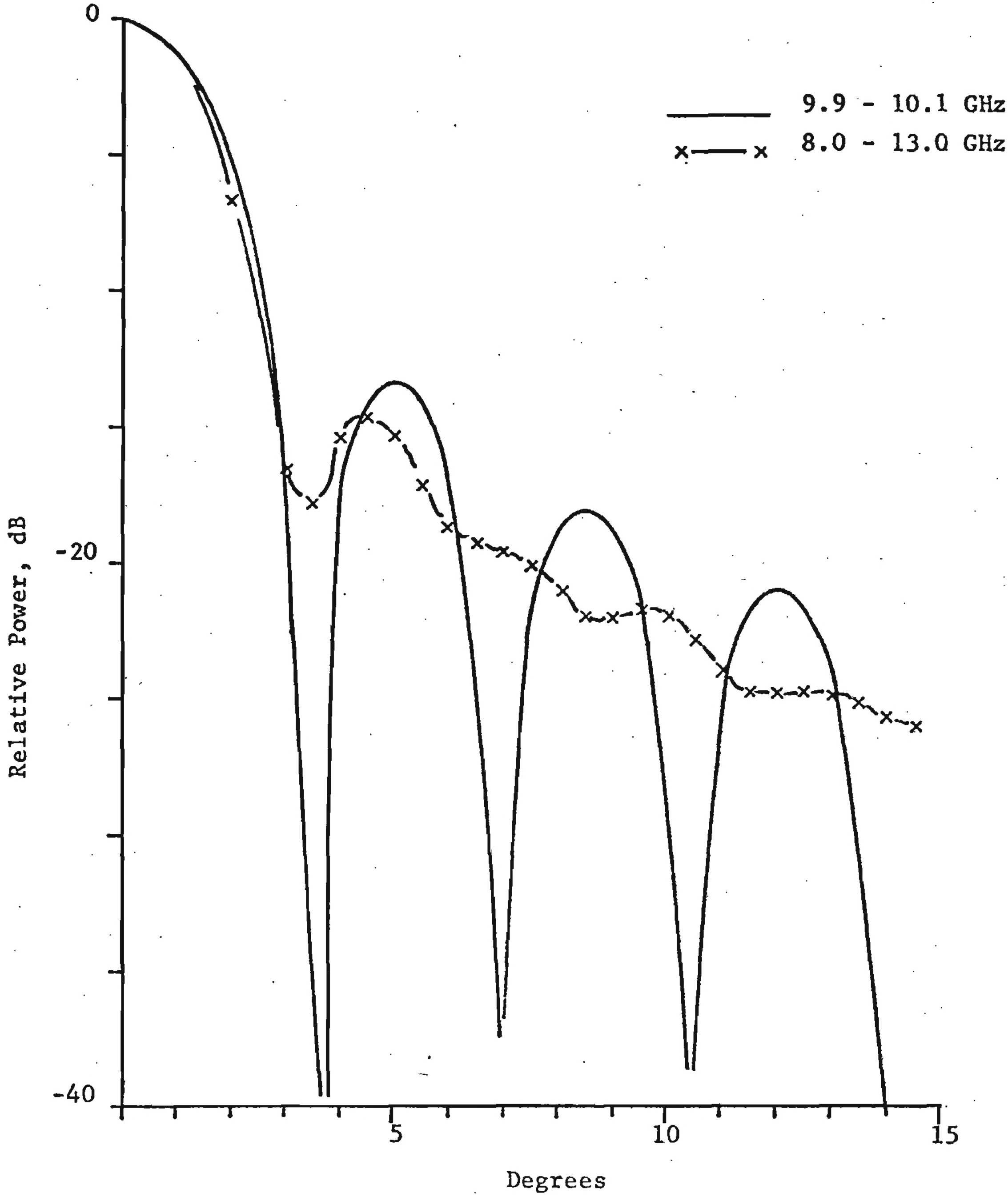


Figure 20. Broadband pattern for uniformly illuminated rectangular aperture, 5 GHz bandwidth

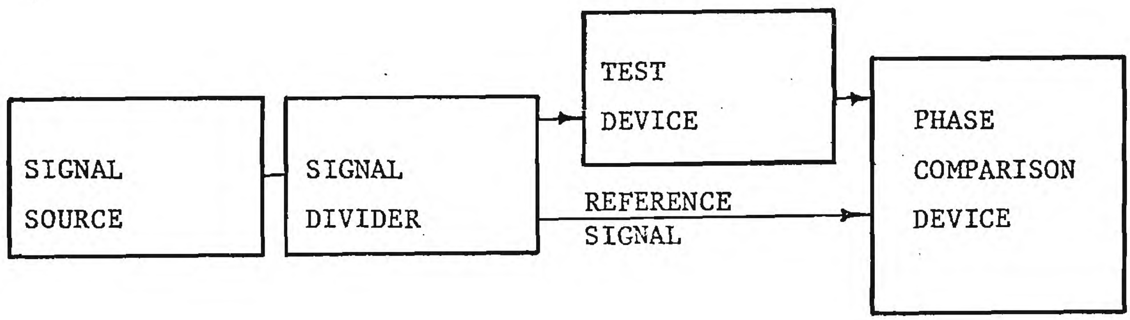
the antenna.

As discussed in Section I-C, two techniques for phase measurement have been considered. Instrumentation of both techniques is feasible with the current state-of-the-art. Adaptation of the conventional separate reference channel technique for sweep frequency applications can be accomplished through modification of the existing Scientific-Atlanta Series 1750 receivers to incorporate phase-locked frequency tracking between the transmitter and the receiver. The group velocity concept can be instrumented through the addition of sideband processing circuits to the modified Series 1750 receiver. Thus, although the instrumentation of a sweep frequency phase/amplitude system is currently feasible, it may be seen from Figures 5 and 6 that a great deal of relatively complex equipment is required. In addition, implementation of a sweep frequency phase/amplitude system requires a system analysis, a definition of computer interface requirements, a software development, an analysis of system calibration requirements, and a definition of system calibration procedures. Thus, a relatively modest expenditure of time and money is required to implement the desired phase/amplitude system. Accordingly, the approach which has been defined through the studies under this contract is based on (1) the completion of the preliminary design of the relatively inexpensive Hybrid system to the component level so that it can be implemented immediately, and (2) the completion of the feasibility investigation of the phase/amplitude system. This second item includes the evaluation of the application of both the separate reference channel direct phase measurement technique and the group velocity measurement technique, to the broadband antenna measurements problem, the investigation of calibration requirements and techniques, the estimation of system accuracy, the preliminary system design (not to the component level), and the definition of critical areas where breadboarding may be desired. The Hybrid system design has been presented in Section III, and the results of the phase/amplitude system investigations are presented in this section.

B. SEPARATE REFERENCE CHANNEL CONCEPT

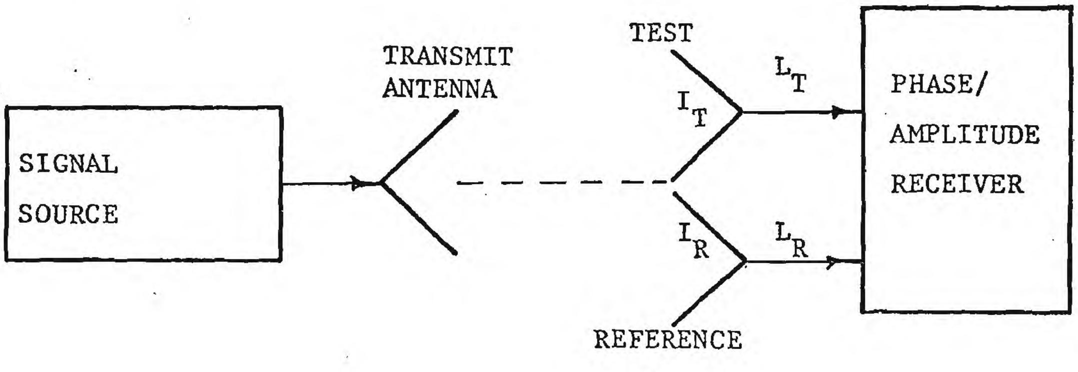
As indicated above, implementation of the separate reference channel phase/amplitude system would employ a modified version of the Scientific-Atlanta Series 1750 receiver. These receivers measure amplitude and phase of a test signal, relative to a reference channel signal. For complete antenna diagnostics, both amplitude and phase information are required. Amplitude versus frequency information which is relative to absolute gain versus frequency can be provided by use of a standard gain antenna in the reference channel. Measurement of the test antenna's phase properties places additional phase calibration requirements on the reference channel. These additional calibration requirements are defined in the following paragraphs.

Conventional laboratory phase measurements may be performed with a setup such as that sketched below.



Phase measurements may be grouped into three categories - insertion, incremental, or comparative. In an insertion phase measurement, the relative phase shift between the test channel and the reference channel is determined both before and after the test device is inserted in the channel. The difference in the two readings is the insertion phase of the device for that particular set of conditions. In incremental measurements, the test device is inserted in the test channel, and relative phase between the test and reference channels is measured as some parameter, e.g. frequency, is changed. Since only incremental phase shift is of interest in this measurement, zero is artibrarily established at some point in the measurement. In a comparative measurement, phase shift of the test device is determined relative to that of a designated standard.

In the antenna applications under consideration here (pulse diagnostics, antenna effect on FM pulse compression, etc.), it is the test antenna's incremental phase shift over a specified frequency range which is of interest. The measurement setup for measuring the incremental phase shift of the test antenna is diagramed below. It will be shown that to determine incremental phase shift of the test antenna versus frequency, incremental phase shift properties of the reference channel must also be known, and any difference between physical path length in the test and reference channels must be accounted for.



The symbols to be used in defining the calibration requirements are illustrated in the above test setup and are defined below:

I_T = insertion phase of test antenna from aperture plane to antenna terminals,

I_R = insertion phase of reference antenna from aperture plane to antenna terminals,

L_T = physical length from test antenna terminals to receiver input, and

L_R = physical length from reference antenna terminals to receiver input.

To measure the variation of I_T with frequency, the calibration requirements for the reference channel must be defined. The absolute value of I_T is not required; that is, the phase measurement of interest here is the manner in which I_T varies as a function of frequency across some specified frequency range.

At some beginning frequency, the measured phase difference between the test channel and the reference channel, $\Delta\Phi_1$, can be expressed as

$$\Delta\Phi_1 = \left[I_T(1) + \frac{2\pi}{\lambda_1} L_T - \Phi R(1)\right] /N, \tag{23}$$

where $\Phi R(1)$ is total phase delay in the reference antenna plus the transmission line L_R, and the integer N accounts for the multiple-of-360^o ambiguity in the measured phase difference. If the frequency is now changed by some amount to ω_2 such that the measured $\Delta\Phi$ does not change by more than 360^o, then,

$$\Delta\Phi_2 = \left[I_T(2) + \frac{2\pi}{\lambda_2} L_T - \Phi R(2)\right] /N. \tag{24}$$

Combining these two expressions yields

$$\Delta\Phi_1 - \Delta\Phi_2 = I_T(1) - I_T(2) + \frac{2\pi L_T}{\lambda_1} - \frac{2\pi L_T}{\lambda_2} - \Phi R(1) + \Phi R(2). \tag{25}$$

By extending the method used to derive this last expression to additional frequencies, it is clear that the change in measured phase difference as a function of frequency change can be expressed as

$$\Delta\Phi(\omega) = \Delta I_T(\omega) + \Delta EL_T(\omega) - \Phi R(\omega) \ . \tag{26}$$

That is, an increase in measured phase difference between the test and reference channels is due to increased phase delay in the test antenna plus an increased electrical length ΔEL_T of the test antenna line link, minus any increased phase delay in the reference antenna channel. As written, this equation is algebraically correct for increased phase delay in any term of the equation. A decreased phase delay would change the sign of that term.

The calibration required to determine $\Delta I_T(\omega)$ from the measured $\Delta\Phi(\omega)$ may be deduced from the last expression above. The change in electrical length of the test antenna transmission line (L_T) must be known as a function of frequency, and the phase characteristics of the reference channel must be known as a function of frequency. Ideally, the transmission line from the test antenna to the receiver input should have a linear phase shift versus frequency property, and the reference antenna-plus-transmission link L_R should be calibrated as a unit. In addition, any difference in L_T and L_R should be small to prevent large excursions in the phase difference between the test and the reference channels as the frequency is changing. Under these conditions, any measured phase dispersion would be the phase dispersion of the test antenna minus the dispersion of the reference channel. Methods of phase calibrating a reference antenna are discussed in Sub-section IV-D.

C. GROUP VELOCITY CONCEPT

Systems first considered for satisfying the program objective involved measuring the phase versus frequency directly; however, these systems required a separate microwave transmission channel to transmit a phase coherent replica of the test signal between the antenna range transmitting and receiving sites. Because the requirement of a phase reference channel may limit the flexibility of these systems, an approach which does not require transmitting a separate replica of the test signal for making phase measurements has also been considered. This approach is based on the concept of group velocity.

Consider the following exponential form for a traveling EM wave

$$\bar{E} = \bar{A}e^{j(\omega t - \beta z)} , \quad (27)$$

where

ω = angular frequency, and

β = phase constant.

For a single frequency, all points of the wave travel, in a homogenous media, with equal phase velocity v_p ($v_p = \omega/\beta$) and the shape of the wave does not change with time. However, to represent an arbitrary periodic time function, an infinite number of monochromatic waves is required. The shape of such a composite wave is no longer preserved in time unless the phase velocity, i.e., the ratio of ω to β, remains the same for all constituent waves of the time function. If this is the situation, the propagating medium is called non-dispersive.

For dispersive media, the phase velocity does depend on ω. Dispersive and non-dispersive media may be further distinguished through the concept of group velocity. Group velocity may be defined as the phase velocity of a wave envelope, i.e., the velocity of a constant phase point on the envelope. Group velocity, u, is related to phase velocity by

$$u = v_p - \lambda \frac{dv_p}{d\lambda} = \frac{d\omega}{d\beta} . \quad (28)$$

For a non-dispersive media, $dv_p/d\lambda = 0$, and the group velocity equals the phase velocity. Since $d\omega/d\beta$ is the slope of the tangent to an ω versus β curve at a point, both $d\omega/d\beta$ and ω/β may be graphically illustrated through an $\omega \cdot \beta$ diagram, such as illustrated in Figure 21. It is clear from the diagram that the slope of a straight line connecting an arbitrary point on the curve with the origin gives the corresponding phase velocity, $v_p = \omega/\beta$. In non-dispersive media, the $\omega \cdot \beta$ curve is a straight line through the origin, and $v_p = u$. However, for the arbitrary curve of Figure 21, the slope of the straight line to the point P is quite different from the slope of the curve at that point.

Thus, recapitulating, in a non-dispersive medium, $v_p = u$ and the shape of a pulse, such as radar pulse travelling through free space, does not change. In a dispersive medium, such as a dispersive delay line used in pulse compression, the shape of the pulse changes. It is desirable, therefore, to determine dispersive properties of all radar system components to ensure that pulse distortions due to unwanted dispersion do not occur. The system under consideration arrives at a measure of dispersion through a measurement of sideband differential phase of a single-tone modulated carrier.

For illustration consider the following expression for a amplitude modulated wave*

$$e(t) = (E_c + m E_c \cos \omega_a t) \cos (\omega_c t + \delta), \tag{29}$$

where $e(t)$ = instantaneous amplitude,

t = time,

m = modulation index ≤ 1,

E_c = peak carrier amplitude,

ω_a = amplitude modulation radian frequency,

ω_c = carrier radian frequency, and

δ = arbitrary phase difference between carrier and modulating signals.

*Other modulation methods which preserve phase coherency of resulting sidebands can be used. AM is used here for simplicity in the resulting equation.

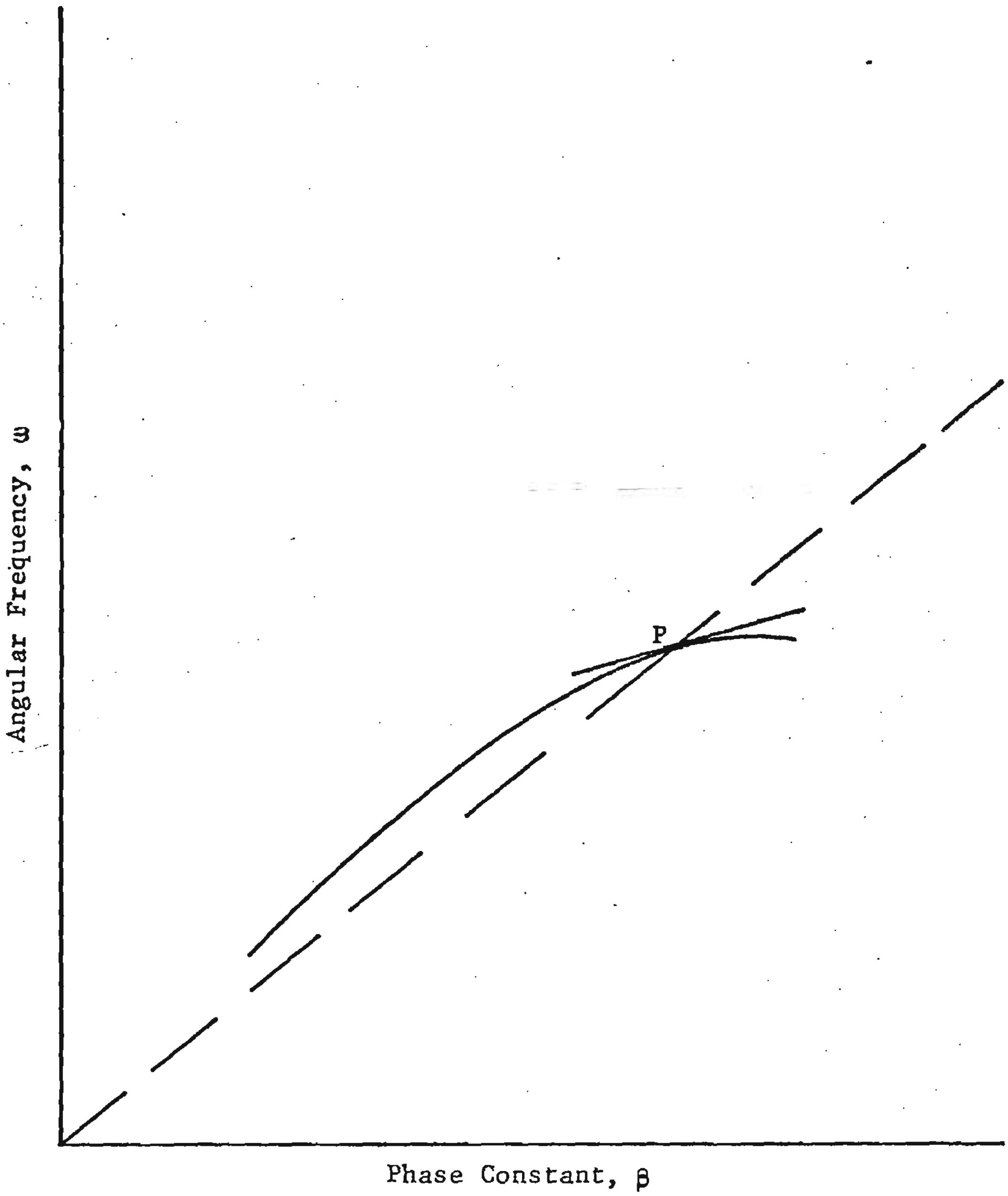


Figure 21. Schematic ω - β′

This equation can be trigonometrically expanded and rewritten as

$$e_c(t) = E_c \cos(\omega_c t + \delta) + \frac{mE_c}{2} \cos(\omega_c t - \omega_a t + \delta) + \frac{mE_c}{2} \cos(\omega_c t + \omega_a t + \delta). \tag{30}$$

The last two terms represent the lower sideband (LSB) and the upper sideband (USB), respectively:

$$LSB = \frac{mE_c}{2} \cos(\omega_c t - \omega_a t + \delta), \text{ and} \tag{31}$$

$$USB = \frac{mE_c}{2} \cos(\omega_c t + \omega_a t + \delta). \tag{32}$$

The instantaneous phase of these sidebands is represented by

$$\Phi(LSB) = \omega_c t - \omega_a t + \delta, \text{ and} \tag{33}$$

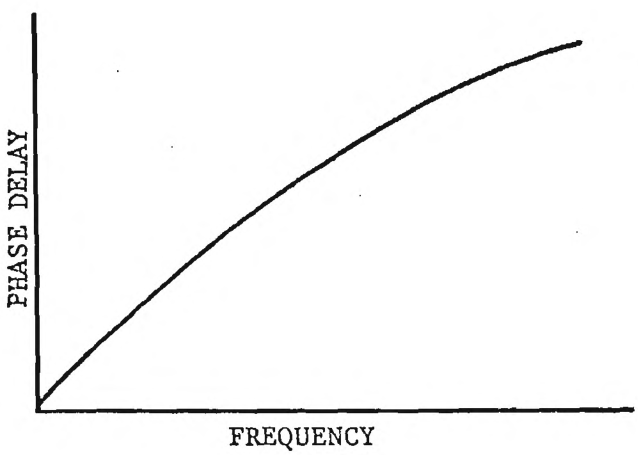
$$\Phi(USB) = \omega_c t + \omega_a t + \delta. \tag{34}$$

In the present application, the test antenna will delay these sideband phases. The phase delay introduced may be expressed as Φ_u and Φ_ℓ for the upper and lower sidebands, respectively. Modifying Equations 33 and 34 to incorporate the phase delays yields

$$\Phi(LSB) = \omega_c t - \omega_a t + \delta + \Phi_\ell, \text{ and} \tag{35}$$

$$\Phi(USB) = \omega_c t - \omega_a t + \delta + \Phi_u. \tag{36}$$

Although sideband phase information may be processed in various ways to obtain a measure of antenna dispersion, it is desired to obtain phase delay versus frequency data such as illustrated below.



The upper and lower sidebands are separated, and a measure of the difference in sideband phase delay $\Delta\varphi_d$ for a given sideband separation $\Delta\omega$ $(\Delta\varphi_d/\Delta\omega)$ is obtained.

This $\Delta\varphi_d/\Delta\omega$ is the average group delay over the range $\Delta\omega$. If the φ_d versus ω curve is linear over $\Delta\omega$ with constant slope $\Delta\varphi_d/\Delta\omega = -\tau$, then the pulse is simply delayed by τ with no distortion. Therefore, the interest lies in measuring the antenna frequency response $\underline{H}(j\omega)$ which represents any deviation of antenna response from constant amplitude and linear phase over the bandwidth of interest (the subscribed bar indicates that $\underline{H}$ is a complex quantity). The conventional phase measurement approach which employs a separate reference channel measures φ versus ω directly, whereas the differential sideband measurement provides $\Delta\varphi/\Delta\omega$ so that an integration is required to obtain $\varphi(\omega)$.

Without a delay-calibrated reference path, one could either attempt to measure $\frac{d^2\varphi}{d\omega^2}$ and integrate twice or to measure $\frac{d\varphi}{d\omega}$ and integrate once to obtain $\varphi(\omega)$. In practice, the first and second derivatives are approximated by first and second differences, designated φ' and φ''. These approximations are defined by

$$\varphi' = \frac{\Delta\varphi}{\Delta\omega}, \text{ and} \tag{37}$$

$$\varphi'' = (\varphi'_1 - \varphi'_2)/\Delta\omega, \tag{38}$$

where φ'_1 and φ'_2 are successive values of φ' taken at intervals of $\Delta\omega$. This approximation is necessary because with the modulation, or sideband, approach a finite sideband separation is required.

Without an independent link to the modulation source, it is possible to measure φ' in a dynamic measuring process in which the frequency band is swept continuously at a linear rate. With a link to the modulation source, it is possible to measure φ' directly by static measurements. Successive measurements at incrementally spaced frequencies over the frequency band are required, but these can be made over arbitrarily selected time increments as long as the modulation reference is sufficiently stable. A link to the source can be obtained by hard-wiring, by a microwave or laser link, or by synchronization of oscillators at the transmitting and receiving sites with WWVB. To achieve the desired accuracy, resolution, and stability, the measuring system proposed herein employs a modulating-tone reference from the source. For relatively short ranges, a coaxial cable transmitting the modulation reference (5-MHz) is a practical approach. This cable does not have to be phase calibrated, but its delay characteristic should be stable over the measurement time. For long ranges or if a direct link is impractical, common synchronization of both the modulating signal at the transmitter and the modulating reference signal at the receiver with WWVB is the preferred alternative.

Because the measured phase delay will represent the total phase delay from the transmitter to the receiver, determining the test antenna phase delay will require calibration of the phase delay due to other elements in the network. These elements include the signal source, the transmission line from the source to the transmitting antenna, the transmitting antenna, the transmission line from the test antenna to the receiver, and a reference antenna. Calibration methods are discussed in the following subsection.

D. CALIBRATION METHODS

Consider the general phase measurement system diagrammed in Figure 22. This figure illustrates all the signal requirements for measurements of antenna phase response either by direct measurement of $\varphi(\omega)$ or by measurement of group delay $\varphi'(\omega)$. The direct phase measurement employs a two-channel phase-amplitude receiver with a reference channel input (channel B) in addition to the test signal input. The signal source is swept continuously across the

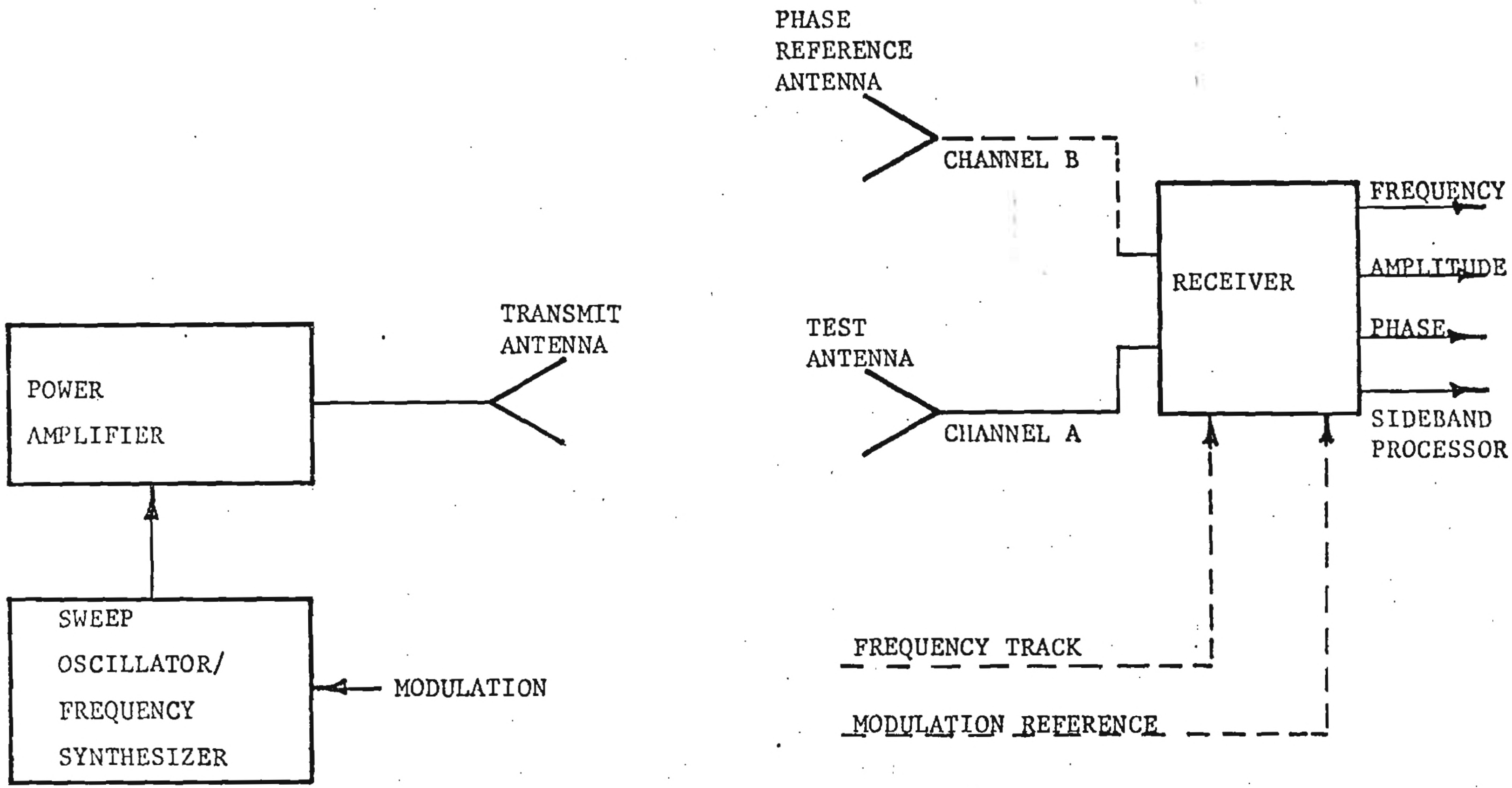


Figure 22. General antenna measurement system

desired frequency range with no modulation. By ensuring an adequate signal to noise ratio, the reference channel input can also be used for frequency tracking. The receiver tracks the sweep signal at a rate up to 10 MHz/msec. No additional tracking or reference inputs to the receiver are required for this type of measurement. The receiver outputs are the test frequency and the amplitude and phase response of the test antenna. The amplitude and phase data are relative to the reference channel response, and they must be corrected by calibration data for the reference antenna.

For group delay measurements, the channel B reference input of the receiver is terminated, the signal source is modulated with a 5-MHz (tentative) modulation tone, and the carrier frequency is stepped sequentially in, for example, 100-MHz steps, and external frequency tracking and modulation reference signals are provided. A separate frequency tracking channel is used since the amplitude of the carrier signal through the test antenna may vary rapidly as the test antenna is rotated. The output of the sideband processor is the sideband differential phase delay.

The measured complex (phase and amplitude) response, $\underline{G}$, of the complete system may be generally represented by

$$\underline{G} = \underline{F}_S \, \underline{H}_C \, \underline{H}_S \, \underline{H}_R \, \underline{H}_D, \tag{39}$$

where

$\underline{F}_S = \underline{F}_S(j\omega)$ is the response of the swept frequency signal source,

$\underline{H}_C = \underline{H}_C(j\omega)$ is the response of the transmission path connecting the source with the transmitting antenna,

$\underline{H}_S = \underline{H}_S(j\omega, \varphi_o, \theta_o)$ is the response of the transmitting antenna at the peak (φ_o, θ_o) of its beam pattern, assumed to be directed toward the receiving antenna(s).

$\underline{H}_R = \underline{H}_R(j\omega, \varphi, \theta)$ is the response of the receiving antenna, and

$\underline{H}_D = \underline{H}_D(j\omega)$ is the response of the transmission path from the plane defining the terminals of the receiving antenna to the terminals at which the output response $\underline{G} = \underline{G}(j\omega, \varphi, \theta)$ is measured.

It may be seen that to obtain the antenna response $\underline{H}_R$ from a single measurement of $\underline{G}$, all other quantities in Equation 39 must be known individually, or their product must be known. For substitution measurements, a calibrated standard antenna must be available whose phase and amplitude properties are known. There are several possible techniques for phase and amplitude calibration of a reference antenna and these are discussed below. The conventional standard gain antenna has been calibrated in amplitude only.

1. Three-Antenna Group Delay Technique

First, a receiver measurement of the source and transmission line responses ($\underline{F}_S$, $\underline{H}_C$, $\underline{H}_D$) must be made. This measurement is made by connecting the source transmission line directly into the receiver connecting line. The impedances of these various components must be matched so that reflections between them are negligible. Calibration of the source-to-receiver response is made by measuring

$$\underline{G}_C = \underline{F}_S \underline{H}_C \underline{H}_D. \tag{40}$$

These factors must now remain unaltered for calibration of the reference antenna.

The three-antenna method requires two "reference" antennas, R and R', and the transmitting antenna. Either R or R' can later serve as a reference antenna. Three successive measurements are made using three different pairs of these antennas. The required setups are illustrated in Figure 23. The following three equations represent the measured responses

$$\underline{G}_1 = \underline{F}_S \underline{H}_C \underline{H}_S \underline{H}_R \underline{H}_D, \tag{41}$$

$$\underline{G}_2 = \underline{F}_S \underline{H}_C \underline{H}_S \underline{H}_R' \underline{H}_D, \text{ and} \tag{42}$$

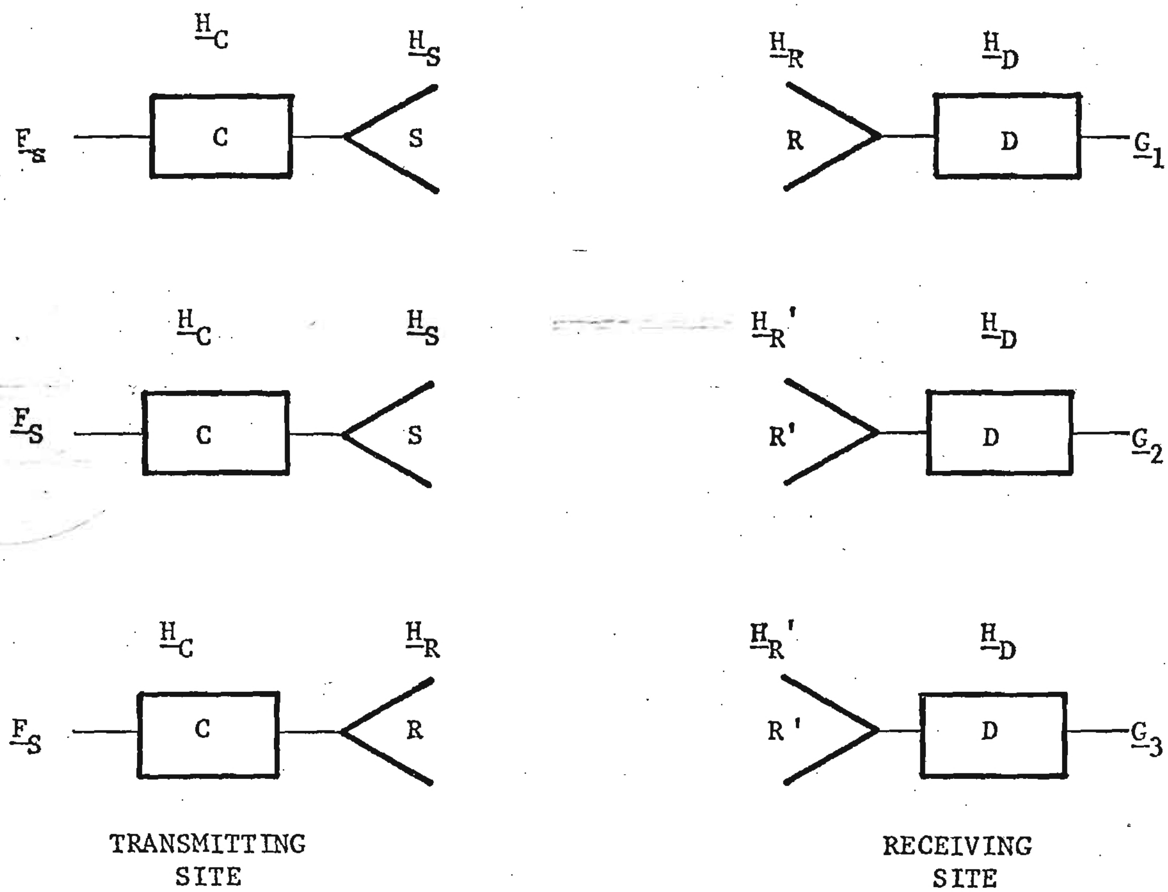


Figure 23. Three antenna group delay calibration method

$$\underline{G}_3 = \underline{F}_S \, \underline{H}_C \, \underline{H}_R \, \underline{H}_R{}' \, \underline{H}_D \, . \tag{43}$$

With $\underline{F}_S \, \underline{H}_C \, \underline{H}_D = \underline{G}_C$ known from the previous measurement, the three antenna responses are found to be

$$\underline{H}_R = \frac{\underline{G}_1 \, \underline{G}_3}{\underline{G}_2 \, \underline{G}_C} \, , \tag{44}$$

$$\underline{H}_R{}' = \frac{\underline{G}_2 \, \underline{G}_3}{\underline{G}_1 \, \underline{G}_C} \, , \text{ and} \tag{45}$$

$$\underline{H}_S = \frac{\underline{G}_1 \, \underline{G}_2}{\underline{G}_3 \, \underline{G}_C} \, . \tag{46}$$

It is seen that antenna calibration depends on four separate measurements, interchanging of antenna pairs from transmitter to receiver, and direct connection of transmitting and receiving hardware at one site. The accuracy of this calibration is critically dependent on maintaining good impedance matches with each test arrangement and on transmitter/receiver stability over the measurement period.

2. Separate Reference Channel Direct Calibration

The two-channel phase/amplitude receiver can be used for calibration of the reference antenna if a suitable link to the signal source can be provided. For electrically small antennas, the far-field requirement can be met at ranges for which direct coaxial cable links are practical. A phase-compensated cable with nearly linear phase delay characteristics is available. It appears practical to use this type of cable on antenna ranges of approximately 100 feet at frequencies up to about 10 GHz. A 100-foot range could be used for S-band reference antennas with gains of about 30 dB or less. At 4 GHz, the loss in 100 feet of this type of cable is about 5 dB. Cable assemblies with phase deviation of a few degrees (<10) across the 2-4 GHz range are available. With this technique, identical antennas are required (see Appendix II) for transmit and receive, either one of which will become the reference antenna. The combined phase delay of the two antennas is measured with the reference signal provided via the coaxial cable. One-half of the measured phase delay is then assigned to each antenna. The calibration accuracy of this technique is limited by the phase stability of coaxial cable and by the requirement of two identical antennas.

3. Antenna Impedance and S-Parameter Method

The antenna impedance and S-parameter method has been used to determine the amplitude and phase properties of broadband antennas over 10 to 1 bandwidths. This method consists of broadband measurements of both the input reflection coefficient of an antenna and the forward transmission coefficient between a transmit/receive pair of identical antennas (see Appendix II). The amplitude and phase properties of a reference antenna can be determined from measurements of both the input reflection coefficient S_{11} and the forward transmission coefficient S_{21} over the band of frequencies for which the antenna will be used as a reference. The fundamental equations which describe the antenna as a transmitter and receiver and the expression for the voltage gain of a system composed of two identical antennas separated by a free-space transmission path of range r are presented in Appendix III.

The first step in this method of calibrating a reference antenna consists of a measurement of the input reflection coefficient of the antenna. The antenna impedance Z_a is related to the reflection coefficient S^o_{11} by

$$S^o_{11} = \frac{Z_a - Z_c}{Z_a + Z_c} , \tag{47}$$

where Z_c is the characteristic impedance of the transmission line. The notation S^o_{11} is adopted to denote that the antenna is radiating into free space. Solving for Z_a yields

$$Z_a = Z_c \frac{1 + S^o_{11}}{1 - S^o_{11}} . \tag{48}$$

A basic setup for measuring the reflection parameter S_{11}^{o} is shown in Figure 24. The line stretcher allows the plane at which the measurements are made to be extended to the terminals of the antenna as well as the equalization of the transmission line lengths between the reference and test channels. If these line lengths are equalized, no differential phase change is observed between the reference and test channels as the frequency is swept over the band.

The next step in calibrating a reference antenna is the determination of the forward voltage gain A_v by measurements of the forward transmission coefficient S_{21} and the input reflection coefficient S_{11}' for an arbitrary load impedance. In most cases, the load impedance Z_L at the output terminals of the receiving antenna will be chosen for a matched condition such that $Z_L = Z_C$ and then $S_{11}' = S_{11}$. For most transmit/receive test configurations of interest, the transmit and receive antennas are separated by sufficiently large distance that the radiation resistance of the transmit antenna is insensitive to the presence of the receive antenna. Thus, the previous measurement of S_{11}^{o} will suffice to determine the input reflection coefficient S_{11} under arbitrary test conditions.

Figure 25 shows the setup for the measurement of the forward transmission coefficient S_{21}. As discussed in Appendix III, the forward voltage gain A_v is related to S_{11} and S_{21} by

$$A_v = \frac{S_{21}\ (1 + \Gamma_\ell)}{(1 - S_{22}\ \Gamma_\ell)\ (1 + S_{11})} , \tag{49}$$

where Γ_ℓ is the reflection coefficient looking out from the receiving antenna terminals. In most practical situations, Γ_ℓ is approximately zero so that

$$A_v = \frac{S_{21}}{1 + S_{11}} . \tag{50}$$

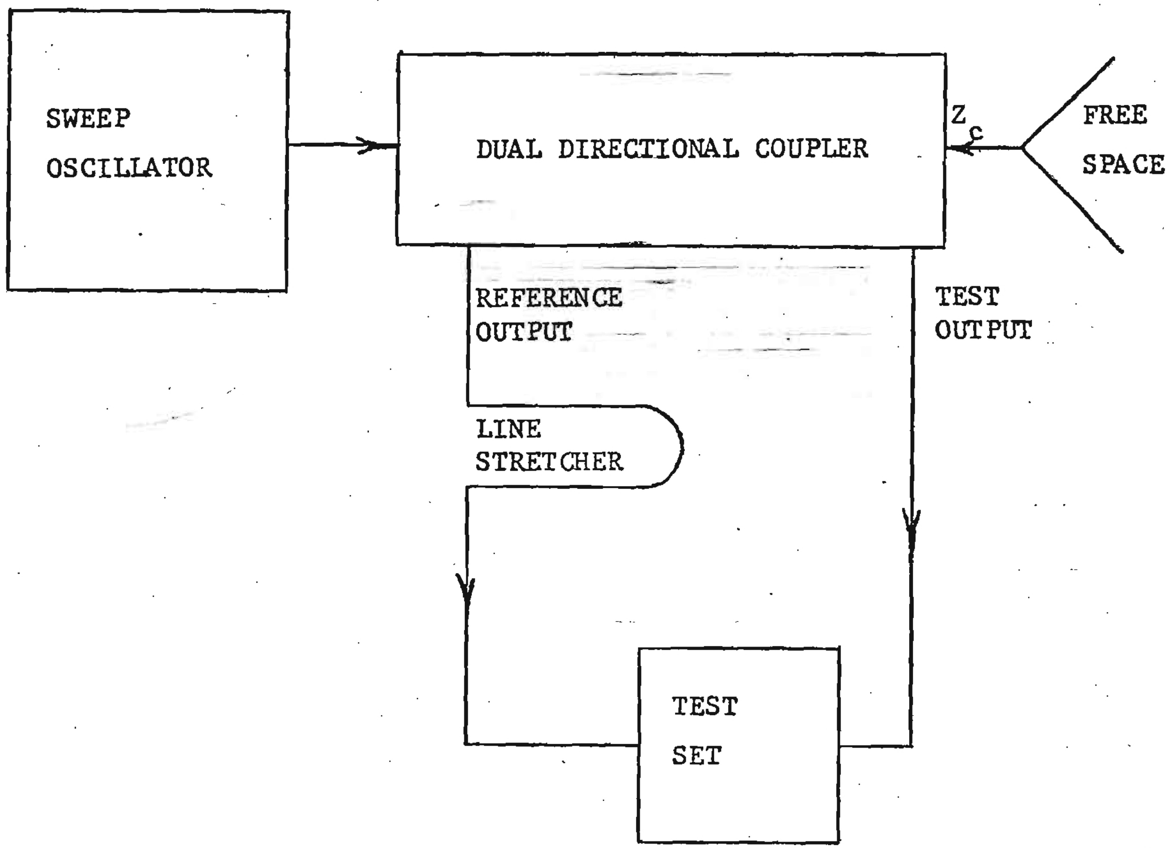


Figure 24. Basic setup for measuring reflection Coefficient S_{11}

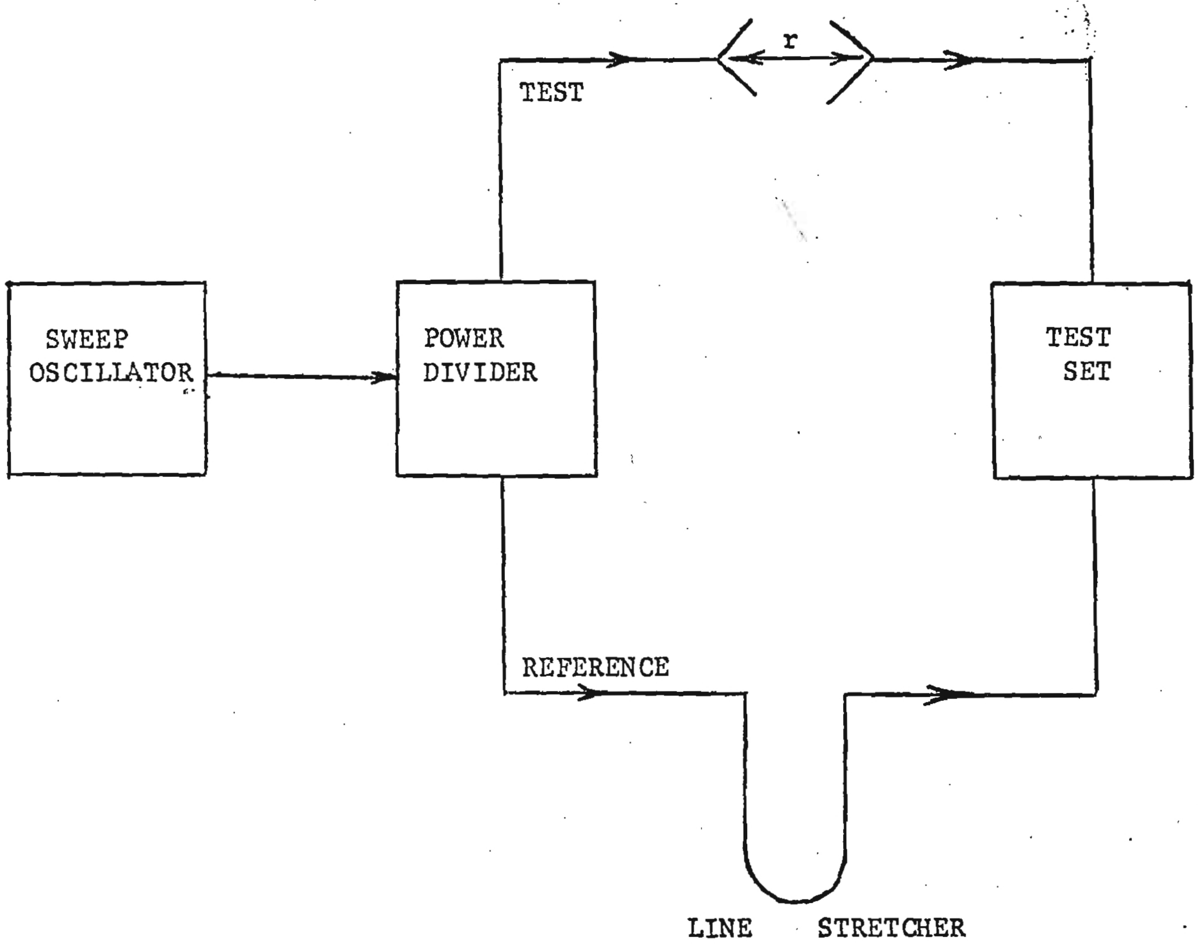


Figure 25. Setup for measuring forward transmission coefficient S_{21}

The line stretcher is used to refer the measurement reference planes to the antenna terminals. The subsequent measurement of S_{21} will then include the phase shift due to the free space range r. If precise range information is available, the effects of the free-space path can be removed by calculation. Also, accurate range information is required to precisely locate the terminal planes of the antenna.

These reflection coefficient and forward transmission coefficient measurements in principal can be performed using standard laboratory test equipment. However, for directive antennas the requirement that a TEM-mode transmission line connect the "identical" transmitting and receiving antennas to the test instrument may present a problem. This requirement may be a problem for directive antennas with far-field requirements since there would be a significant cable loss which would degrade the accuracy of the measurement. Thus, the limitations of this approach appear similar to those of the direct phase comparison approach described in Subsection D-2 above.

4. Free-Space Scattering Method

A free-space method for the determination of an antenna's complex gain (amplitude and phase) versus frequency characteristics exploits the fact that, if the antenna terminals are mismatched, the echo area of an antenna is proportional to the square of the antenna's power gain. Or stated another way, the field scattered from the antenna is proportional to its voltage gain squared. If the magnitude and phase of this scattered field are measured, the complex voltage gain can be determined. (The required equations are developed in Appendix IV.) In developing the method, the amplitude and phase of the scattered field of a sphere of known size is first measured. Next, the amplitude and phase of the scattered field of the antenna for a short-circuit terminal condition is measured, relative to the scattered fields of the sphere.

Finally, the absolute magnitude and phase of the antenna scattered field are determined using the known Mie series solution for the scattered field of the sphere.

A potential difficulty in applying this technique to broadband sweep-frequency measurements is that the background noise cannot be cancelled using a nulling loop as is done in single frequency applications. However, if both the sphere return and the antenna return are much larger than the background noise, the error introduced by the background will be small. Also, it is possible to record the background noise as a function of frequency and then subtract this recorded background noise from the signal-plus-background noise. This method has been successfully employed at Georgia Tech for near-field radar cross section measurements [3] at a single frequency, and it has been determined that background-to-target signal ratios on the order of -3 dB can be tolerated. Experiments have shown that the background signals in an anechoic chamber are reasonably stationary with respect to time over periods of a few hours. Thus, it appears that this "computer nulling" method can be employed in a sweep frequency measurement.

An advantage of this method is that the transmission paths are common for both the sphere and the antenna under calibration. Thus, no error is introduced by the transmission paths. Also, the phase shift from the antenna aperture to the output terminals is directly measured with respect to the sphere. Since the sphere center and the antenna aperture can be optically aligned, a precise measurement of the range from the transmitter to the target is not necessary. This method thus appears to have good potential for determining reference antenna amplitude/phase versus frequency characteristics. Experimental evaluation of this method is desirable.

5. Comparison of Calibration Methods

Standard gain antennas have long been used for the absolute gain calibration of an unknown antenna in amplitude only. However, a reference antenna calibration including both amplitude and phase antenna measurements is more difficult since both the voltage gain and the phase shift versus frequency of the reference antenna must be determined. It should be noted that, to date, no experience in characterizing a reference antenna in both magnitude and phase over a wide frequency bandwidth has been obtained. Thus, the magnitude of the practicable measurement difficulties which may arise can not currently be pinpointed.

The optimum calibration method will depend on the electrical size of the intended reference antenna and on the required accuracy of calibration. For relatively small antennas, direct phase measurements are practical, and they should provide accuracy within approximately 10 degrees. Continuous broadband calibration will require a tracking broadband receiver. However, static measurements at selected frequencies across the desired calibration range can be made with currently available instrumentation.

The three-antenna group delay method is attractive because it can be used with any reasonably sized reference antenna, and identical antennas are not required. Assuming availability of a group delay measuring receiver, each individual measurement for this method can be made with high accuracy. Experimental tests are required to firmly establish the overall calibration accuracy which can be achieved.

The free-space antenna scattering concept is currently limited in application to antennas of relatively small physical size. The required measurements are conveniently made with proper instrumentation. Experimental evaluation of achievable accuracy is desirable.

The antenna impedance and S-parameter method is probably better suited to less directive low frequency antennas.

E. GENERAL HARDWARE REQUIREMENTS

During the investigations into the capabilities and requirements of direct-phase sweep-frequency measurement techniques and of group delay measurement techniques, the two concepts were analyzed separately. However, since the group delay measurement would require a basic receiver plus the sideband processor, this concept separation is not desirable from a hardware implementation point of view. Hardware for implementing both techniques will be discussed.

As discussed in the previous portions of this section, a sweep-frequency direct-phase measurement technique can be implemented by modification of the existing Scientific-Atlanta Series 1750 Phase/Amplitude receiver. The group delay measurement capability can be implemented by including the sideband processing circuits in the modified receiver. A block diagram of the receiver is shown in Figure 26.

The receiver is a phase-locked superheterodyne type which employs harmonic mixing. It is generally similar to the Scientific-Atlanta Series 1750 Receiver, except that it employs a frequency-agile local oscillator and has additional circuits which give it the capability of making group-delay measurements in addition to direct phase/amplitude measurements. All readouts from the receiver are available in digital form by routing the 1-kHz output signals to integral phase and amplitude display units for processing and conversion to digital form. Also shown in Figure 26 are the local-oscillator and phase-lock circuits and the output to the sideband processor, which is shown in Figure 27.

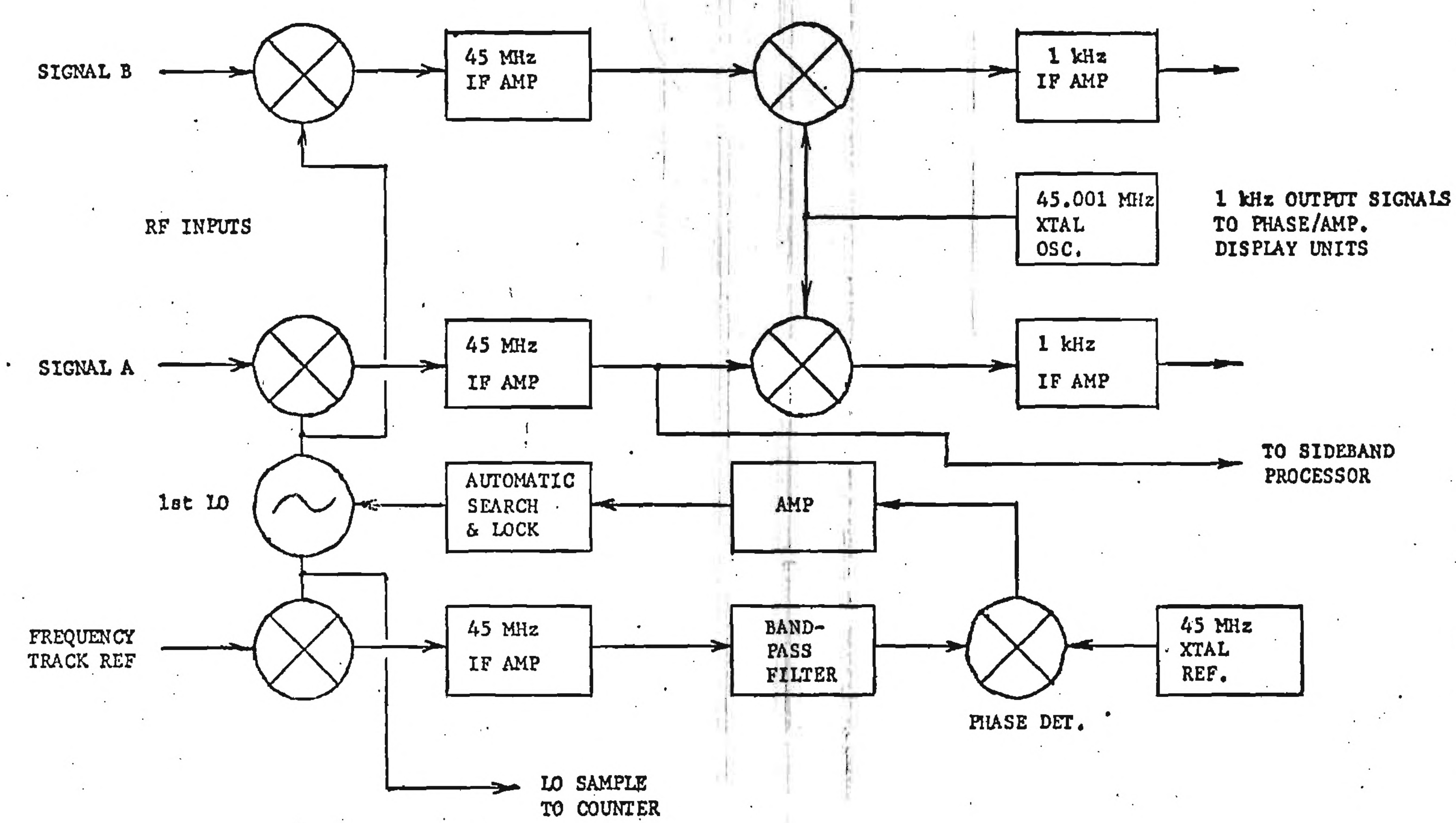


Figure 26. Tracking phase-locked receiver

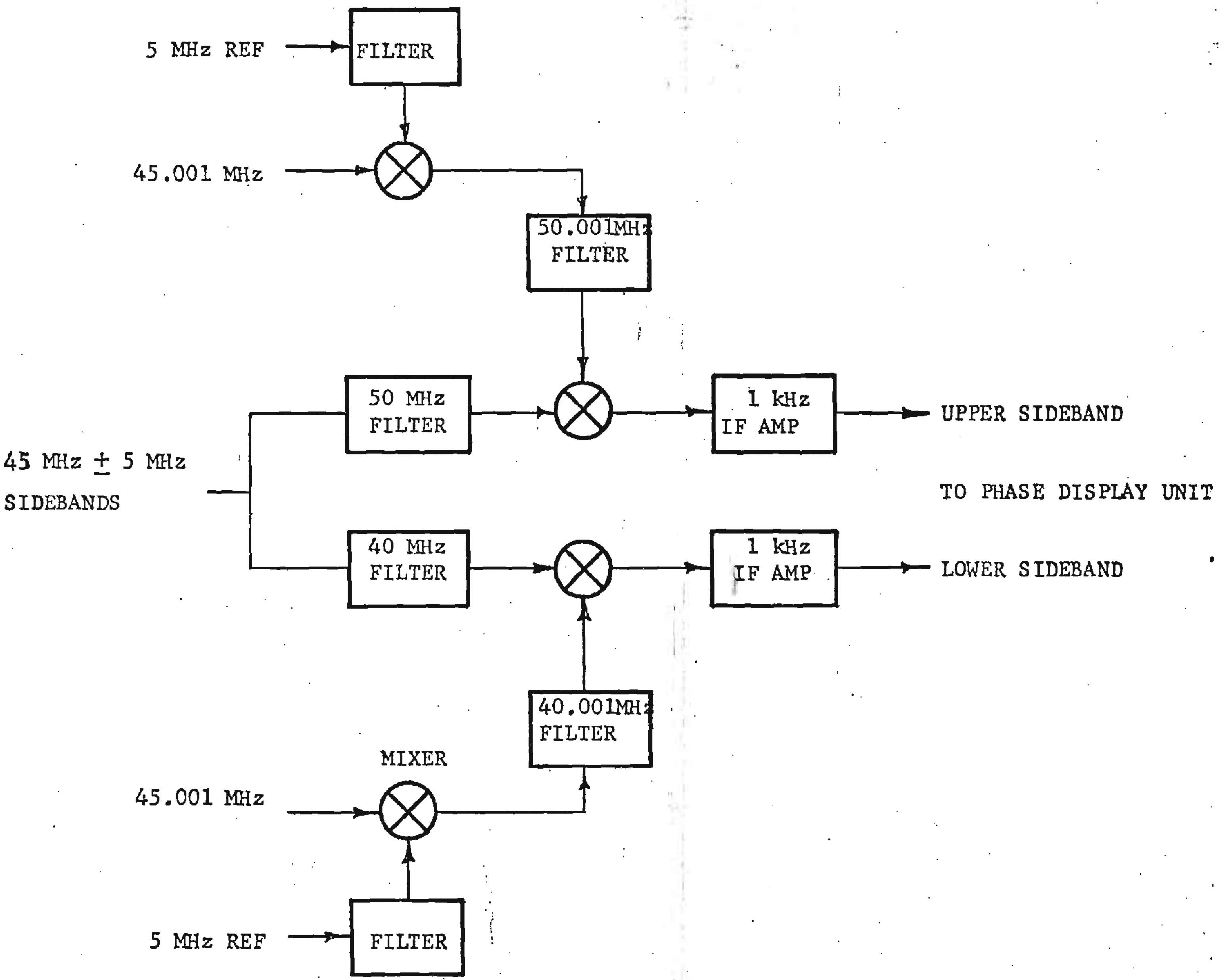


Figure 27. Sideband differential-phase processor

The sideband processor is required for measurement of $\Delta\varphi/\Delta\omega$. The result of the various frequency translations is to provide output signals $(\varphi_1 - \varphi_2 - 2\alpha)|_{\omega_1}$, where ω_1 represents the set of measuring frequencies separated by $\Delta\omega$. The output signals are processed to give the phase difference

$$\Delta\Phi = (\varphi_1 - \varphi_2 - 2\alpha)|_{\omega_1} = \Delta\varphi - 2\alpha|_{\omega_1} , \qquad (51)$$

where α = phase delay in the modulation reference. To avoid error, the 2α term must remain constant during the measuring interval. If this term is constant but non-zero, digital integration of $\Delta\Phi/\Delta\omega$ gives rise to a term which represents a time shift, which is of no consequence. For convenience, α can be varied in setting up the system.

The phase/amplitude measurement system is shown in Figure 28. As discussed above, the sideband processing circuits are incorporated in the receiver, and the transmitter contains the modulator required for generating the sidebands. For clarity, A/D converters are specifically indicated in the receiver output channels. However, the Scientific-Atlanta hardware contains its own digital display and output capabilities. (A multiplexer is indicated for ease in interfacing these outputs with the computer.) The receiver's digital phase unit, whose dynamic range at the input is at least 60 dB, converts the measured phase between the test signal channel and the reference signal (or between the two sidebands) to a digital signal.

The receiver's digital amplitude unit converts the analog amplitude signals from the basic receiver to a digital dB format. It also provides the ratio, in dB, of the inputs from the test and the reference channels, and it can be externally triggered for updating. The minimum dynamic range of this unit is also 60 dB.

The computer-to-pattern-recorder interface (D/A converter), routes selected pattern data in decibel form from the computer to the pattern recorder

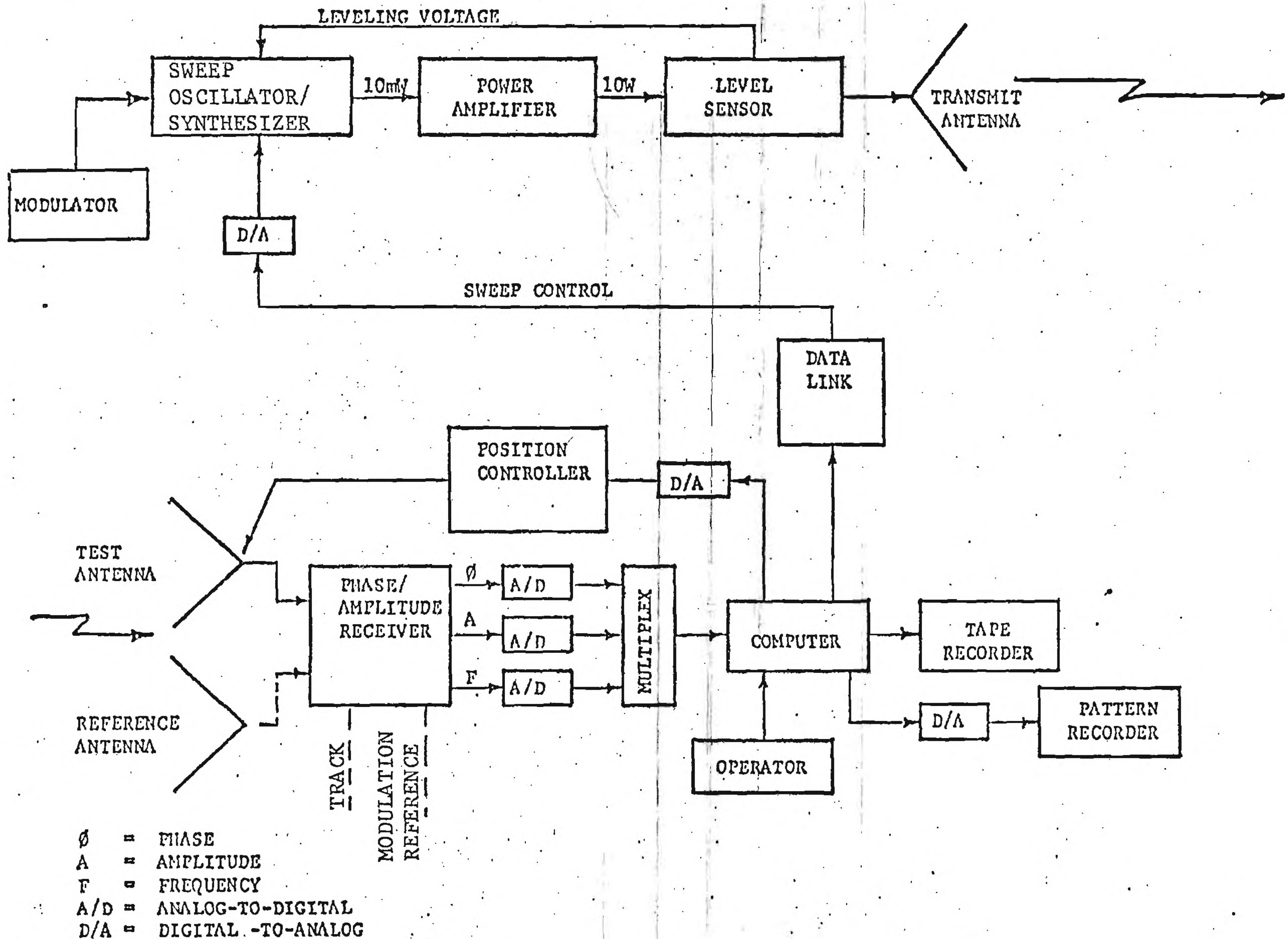


Figure 28. Phase-and-amplitude measurement system

for display. The nominal range of this signal is four decades. Resolution requirements can be met by a 10-bit D/A converter (provides resolution here of about 0.04 dB). Since only one selected frequency component is expected to be recorded for display, the D/A conversion speed is not a critical requirement, and this interface presents no hardware problem.

The computer-to-position-controller interface passes such data as the test antenna start and stop position and the desired pattern cut. Since present positioners cannot be stepped to less than 0.1 degree, extreme resolution is not required in this channel. A resolution of 0.1 degree out of 360 degrees (equivalent to positioner step resolution) can be provided by a 12-bit D/A converter.

Flexibility in the computer-to-transmitter interface which will allow specifying data recording modes with varying degrees of accuracy and with various data rates is desirable. For a "quick look" capability, a simple START SWEEP command which causes the transmitter to linearly frequency sweep at a high rate across its range setting is desirable. The frequency is swept at a constant rate, and the frequency at each sample point is determined by a counter in the receiver. In this mode, the accuracy to which data are obtainable is limited by the uncertainties imposed by the requirement for a finite measurement and conversion time. For more precise data, a slower data rate is required. At some spatial antenna positions, such as near a sidelobe null, frequency sweeping may cause the antenna phase response to undergo rapid changes. Since the receiver output circuits are relatively narrowband, the phase information might be significantly "smoothed." In such a situation, more precise information can be obtained by stepping the frequency in discrete steps with 20 to 50 msec between steps. This 20- to 50-msec time interval would allow the phase output circuit to stabilize before a reading is taken, and the frequency would always be known within its step accuracy. This latter mode of operation can be implemented with a programmable transmitter which responds to the STEP SCAN command from the computer. In addition, for group delay measurements, discrete frequency steps are required with a highly stable frequency during the measurement interval. Data recording rates and sampling requirements are further discussed in the following subsection.

Computer requirements for this system can be met by the Raytheon 703 which has been described in the Statement of Work for this investigation. Based on experience in operating the computer-controlled near-field range at Georgia Tech, software instructions for computer control during data recording are expected to require approximately 4k of storage. Recording phase, amplitude, and frequency at 100 frequency points per spatial position requires 300 data words. Since the Raytheon 703 has a 32k-word memory, phase, amplitude and frequency data for approximately 90 spatial positions could be recorded in the computer memory. The Raytheon 703 has two types of input/output data communication channels, the Direct Input/Output (DIO) and the Direct Memory Access (DMA). Using DMA, 16-bit data words may be interchanged between 703 memory and up to six external devices simultaneously, interlaced with computation/executions. With the 703's basic memory cycle time of 1.75 μsec, data can be accepted at a faster rate than would typically be required (see the next subsection). The 703 computer can also be used to perform some on-site diagnostics; this application will be discussed in a following subsection.

The Statement of Work also specifies that the Hewlett-Packard magnetic tape recorder model 2020E be used for data recording. This is a 7-track machine (6 data bits plus parity) which records at 45 inch/sec and at a byte (6 data bits) density of 556 per inch. Two bytes of information will, therefore, be required to record data in more than six-bit words. Tapes made on this machine can be processed readily by the Honeywell 600-line computers and their associated tape handling equipment. Under proper software control, this recorder can interface with the 703 computer for data storage, and it is suitable for this application.

F. SYSTEM OPERATION

Before describing a typical measurement sequence, guidelines for selecting sampling intervals and data rates will be presented. Since the sweep-frequency direct-phase measurements and the static group delay measurements have somewhat different sampling requirements they will be discussed separately. For sweep frequency antenna measurements, there are two sample variables to be considered. These variables are spatial coordinates and frequency. Basically, there are two sampling approaches which should be considered. With the first approach, the antenna under test would be slowly and continuously rotated. At specified time intervals, dependent on the speed of antenna rotation and the desired spatial resolution, the test signal would be rapidly swept through the frequency range of interest. The frequency sweep rate should be such that the antenna could be considered essentially stationary over the sweep time. On the second approach, the test antenna would be incrementally stepped to new positions and remain stationary while the test signal is swept over the desired frequency range. The relative desirability of the two approaches depends on the test antenna size and the required accuracy of the data. Since test signal sampling is required for either of these approaches, this sampling requirement will be discussed first.

The upper limit of the frequency track rate for the projected phase/amplitude receiver is 10 MHz/msec. If samples every 100 MHz are needed, this would require a sample rate of 1/10 msec, or 100 Hz. Recording phase, amplitude, and frequency increase the total data rate to 300 Hz. This 100 Hz per data channel is compatible with the maximum data rate of the receiver output channels. This data rate also is easily accommodated by the 1.75 μsec memory cycle time of the Raytheon 703 minicomputer. Alternately, at this data rate, the data could be recorded directly on magnetic tape. A magnetic tape recording rate of 15 kHz at 12 bits is typical. Sampling every 100 MHz over a 1-GHz range results in 10 frequency sample points. Recording frequency, phase, and amplitude at each sample point would require storing 30 data words for each frequency sweep. (With proper software "bookkeeping," spatial angle need be recorded only at the beginning of each frequency sweep.)

The most appropriate spatial sampling interval will depend on the test antenna size and the desired accuracy. For illustration, consider an antenna having a 3-dB beamwidth of one degree. (A 1-degree 3-dB beamwidth requires approximately a 25-foot diameter aperture at the S-band frequency of 3 GHz.) Assume that it is desired to record pattern cuts over ± 180 degrees. The optimum positioner scan rate is primarily dependent on two competing factors. These are the desirability of minimizing the total pattern recording time (fast scan) and of minimizing movement of the test antenna during a frequency sweep (slow scan). Examination of these factors shows that a relatively slow antenna scan rate on the order of 0.15 deg/sec is appropriate. Thus, 40 minutes are required to record the entire ± 180 degree pattern. With the previously considered transmitter sweep rate (10 MHz/msec), 0.1 second is required to sweep the full 1-GHz frequency range. During this time, the antenna would sweep through 0.015 degree. This much movement during a frequency sweep would amount to only one one-hundredth of the one-degree beamwidth, and the resulting data uncertainly would be tolerable. Additionally, 0.015 degree is comparable to the readout accuracy of typical positioners.

For a 1-degree total half-power beamwidth, the first nulls would occur near approximately $\pm$ 1.3 degree off beam center, and successive nulls would occur at further 1-degree increments. If 10 sample points per sidelobe provides sufficient resolution, a spatial sample is required each 0.1 degree; because the null-to-null width of the main beam is about 2 degrees, 20 sample points to define the main beam are necessary. At the previously assumed positioner rotation rate (0.15 deg/sec), a new frequency sweep must begin about every 0.67 second in order to sample every 0.1-degree spatial increment. Since only 0.1 second is required per frequency sweep, the frequency sweep rate, spatial sample rate, and positioner scan rate are all compatible. There would be approximately one-half second of deadtime between each data recording interval. For small antennas which can be conveniently step scanned and stopped for data recording, the recording time per pattern may perhaps be reduced to less than 40 minutes. A programmable positioner controller is available which can provide scan increments as small as 0.1 degree (S/A Model 2004 Plus Model 4561 Servo Repeater-Converter). Based on antenna size (and beamwidths) and type of positioner available, the stepped-scan approach should be examined for specific applications. In addition, it may be necessary to use a stepped scan and a slower frequency sweep near null positions where phase and/or amplitude are rapidly changing.

With the group-delay measurement approach, a frequency-stepped approach is required. With the tentatively selected 5-MHz transmitter modulation tone, the sideband separation will be 10 MHz. If the carrier is stepped at 100-MHz increments, essentially the same sampling considerations apply as considered above. That is, in the frequency domain, 10 sample points are produced for each 1 GHz of frequency test range. At each of the 10 points, amplitude, $\Delta\varphi/\Delta\omega$, and carrier frequency are measured for a total of 30 data words at each spatial point. The minimum time between frequency steps should be 10 msec because of the bandwidth of the data channels. These guidelines are seen to be similar to those given for the sweep-frequency approach.

There is an additional consideration in antenna rotation rate which should be noted. A Doppler shift can occur due to antenna rotation during the measurement interval (discussed later under system accuracy). If this occurs, it causes the IF center frequency to deviate from its nominal value of 1 kHz and can lead to a measurement error if the IF shift is too great. If required, however, the antenna can be stepped to avoid any Doppler shift during measurement.

With a calibrated reference antenna available, amplitude and phase measurements can be made directly with the amplitude and phase measuring circuits of the receiver which has been described in Subsection E. Phase can also be measured by using the sideband processor to obtain the differential phase φ' and then integrating. For either case, the phase reference or frequency tracking antenna should be located in relatively close priximity but with negligible coupling to the antenna under test. The total electrical path length from the transmitting antenna through either reference antenna should be essentially equal to that through the antenna under test to prevent large phase excursions of the output with frequency.

Group delay measurements are made by first mounting the calibrated reference antenna on the test mount and taking a set of measurements $\underline{G}_R(\omega)$. Then, a set of measurements $\underline{G}_T(\omega)$ is taken with the test antenna in place. The test antenna transfer properties $\underline{H}_T(\omega)$ are then given by

$$\underline{H}_T(\omega) = \frac{\underline{G}_T(\omega)}{\underline{G}_R(\omega)} \underline{H}_R(\omega) , \tag{52}$$

where $\underline{H}_R(\omega)$ is the calibrated reference antenna characteristics. In a typical data recording sequence, the operator enters the following parameters into the computer: 1) spatial antenna pattern cut, 2) data to be recorded (phase and/or amplitude), 3) antenna rotation rate or step increment, 4) spatial sample increments, 5) frequency sweep or step rate, 6) frequency increments, 7) direct phase or group delay measurements, and 8) frequency component to be plotted in real time on the pattern recorder. The computer generates initial antenna position data which are fed through the D/A converter to the controller which initiates the antenna movement. As the antenna rotates, position data are continuously received by the computer from the encoders on the positioner. As each of the pre-set angular increments is reached, the computer generates a sweep or step frequency command which is transmitted via data link to the transmitter site. The transmitter responds to this command by beginning a frequency sweep over prescribed limits or by sequentially stepping frequency across the measurement range. The receiver is continuously generating phase, frequency, and amplitude data which are fed to the A/D processing units.

As each of the prescribed frequency sample points are reached, the computer reads the phase, amplitude, and frequency data into the computer memory. Antenna position data are recorded before each frequency sweep. The antenna continues to scan or it is stepped after each complete frequency traversal, and the computer repeats the data recording process each time a predetermined angle is reached. In this way, a complete pattern cut with the broadband frequency response at prescribed spatial points is stored in the computer. At convenient intervals (such as the end of each pattern cut, or completion of a frequency sweep), the data stored in computer memory are dumped onto magnetic tape. For example, with the previously considered 30 data words per spatial point and with approximately 28K of Raytheon 703 computer memory for data recording, data for about 900 spatial positions could be stored in computer memory. If less than a 360° pattern or less than 0.1 degree resolution is sufficient, perhaps a full spatial pattern cut can be recorded before memory dump. The computer can select data at any specified frequency for plotting of a "real time" pattern. The amplitude data at each spatial point may be normalized to the maximum amplitude and converted to decibel form before routing to the recorder.

Although antenna diagnostics which require handling a large amount of data will be performed by a large general purpose off-site digital computer, some on-site diagnostics are practical. Tapes recorded on the HP 2020E digital tape recorder can be read directly by most computers, including the Honeywell 600-line, which would be suitable for off-site diagnostics. All of the data which could possibly be recorded for a single antenna probably will not be used for any one given diagnostic. For example, if phase and amplitude were recorded each 100 MHz over a 1 GHz range (10 x 2 data points) at each 0.1 degree over a full 360 degree sector, this would require recording 72,000 data words per pattern cut. At 14 bits per word, this is approximately 1 million bits of information per pattern cut. All of this information is available for processing if desired, but most diagnostics will not require all of this data. To obtain an average pattern response, only the amplitude data are required. Furthermore, in many cases, average response over angular sectors smaller than the full 360° sector will be of interest. In the case of phase response and time domain pulse diagnostics, many applications will be concerned with situations in which the antenna is pointed directly at the source of radar return. In such cases, only the main beam data are required, which greatly reduces the amount of data which must be handled. In summary, even though <u>all</u> the antenna information is available, sorting only that data which are necessary for a given diagnostic will serve to control computer processing time and data storage requirements.

For on-site diagnostics, the Raytheon 703 can generate average antenna response at each spatial position since this requires handling only the amplitude samples over the desired bandwidth. Use of Fast Fourier Transform techniques may permit some pulse diagnostics as well.

G. MEASUREMENT SYSTEM ACCURACY

The accuracy of the measurement system is influenced by a number of factors, some of which are listed below and discussed briefly in the following paragraphs. The measurement system accuracy will depend, of course, on the final design and construction details as well as operational factors. Therefore, this subsection contains only a discussion of the major contributors to error with some general statements as to the accuracy that is achievable.

The measurement system accuracy is dependent on the following factors:

1. Reference antenna calibration accuracy,
2. Impedance,
3. Receiver measurement accuracy,
4. System stability and repeatability, including modulation reference system stability,
5. Dynamic errors,
6. Random noise and signal/noise ratio, and
7. Atmospheric propagation anomalies.

Several methods for calibrating the reference antenna have been discussed. Since the test antenna measurement accuracy will depend on the reference antenna calibration accuracy, to obtain highly accurate test results, a highly accurate calibration method must be used. Assuming availability of a high accuracy measurement system, the three-antenna calibration method is potentially very accurate. A source of error in this method is drifts and instabilities while the three antennas are being interchanged for measurements. Recall also that the transmitter must be transported to the receiver site and connected directly to the receiver for one of the measurements required by the three-antenna method. To obtain high accuracy, not only must the measurement system characteristics be repeatable for both locations of the transmitter, differences due to movement or substitution of RF cables must be avoided. The free space scattering method also is capable of high accuracy with relatively small antennas. Both the direct phase comparison and the antenna impedance/S-parameter methods are expected to be less accurate.

Mismatch errors during tests can be caused by reflections at the various interconnections. The magnitude of amplitude losses and phase uncertainties can be calculated if the reflections are known. Since the measurement system must operate over a broad frequency range, it will probably be necessary to use precision attenuators or broadband isolators to avoid multiple reflections.

The IF and the amplitude/phase measurement units of a modified Series 1750 receiver will be similar to those of the standard Scientific-Atlanta Series 1750 Phase-Amplitude Receiving System. With the Series 1750 system, high-accuracy measurements can be made over a wide dynamic range of the input signal level. Typical accuracies are as follows:

Dynamic Range	Phase Accuracy	Amplitude Accuracy
0-10 dB	0.25°	0.1 dB
0-20 dB	0.5°	0.2 dB
0-40 dB	2°	0.3 dB
0-60 dB	5°	1.0 dB

The Series 1750 system provides a phase resolution of 0.1°, and an amplitude resolution of 0.01 dB over a 20 dB range and of 0.1 dB over a 60 dB range. Further, because a modified Series 1750 is compatible with the group delay measuring system, the accuracy for this system should also be comparable to that shown above. To achieve these results, however, care must be exercised in the design and implementation to prevent errors from RF or IF leakage between channels, pickup of internal reference signals, amplifier non-linearities etc. Particular care must be exercised with the RF circuitry to minimize error due to cross talk between channels. Excessive rates of amplitude and phase variation must be avoided during measurements.

For group delay measurements in which standard and test antennas are interchanged, the system must have good long-term stability in order not to require frequent recalibration of the system. Based on experience with the Series 1750 system, phase and amplitude stability of 0.5° and 0.2 dB over a 20 dB range for a period of one to four hours appears reasonable for operation at a relative constant temperature after warm-up. For direct phase measurements with a calibrated antenna in the reference channel, system stability is not critical.

For extreme accuracy over limited angular segments, dynamic errors caused by scanning frequency while continuously rotating the antenna can be avoided. The frequency can be stepped and the dwell time adjusted as necessary to provide the maximum accuracy.

As indicated earlier, a frequency shift will occur between the signal channel input and the reference or frequency track channel while scanning in frequency, if there is a difference in RF or local oscillator path lengths between the channels, or if the antenna under test is not rotated about its phase center. For maximum accuracy, when these conditions exist the frequency shift should be limited to a few Hertz by limiting the rotation rate.

When making direct phase measurements with a reference antenna employed to receive a reference signal at the test frequency, atmospheric disturbances could cause a random phase variation of the reference signal relative to the signal from the test antenna. However, under normal weather conditions, if the reference and test antennas are separated by not more than 100 feet, which normally is the situation, and if the frequency is in the 1-18 GHz range, the probable random phase variation would be less than one degree [4].

H. LARGE REFLECTOR APPLICATIONS

In many cases, it may not be practical to perform broadband measurements on an entire antenna system. For example, the antenna's electrical size may require a far-field range which is much greater than that which is available or practical. In other cases, the antenna's structure may be such that assembling the antenna on a test range would be impractical. In cases where the entire antenna system either cannot be directly measured or it would be extremely expensive, an alternate approach is desired. There are basically two different types of information which might be required. These are (1) far-field pattern information, and (2) peak gain and insertion phase as a function of frequency. The desired patterns may be either broadband or CW. Since CW information is a special case of the general broadband data, approaches for obtaining the broadband pattern data and the peak gain/phase data are required. It is realized that the phase and amplitude system which is described in this section may offer a possible solution. This approach for obtaining far-field parameters consists of the following: (1) measure broadband phase and amplitude data at the feed terminals with the system which has been described in this section, (2) use this measured feed data to compute aperture phase and amplitude distributions, and (3) transform the aperture distributions to obtain the desired far-field parameters.

In a typical feed/reflector antenna system, the phase center of the feed is not constant over a broad frequency range. Consequently, the phase of the feed radiation pattern depends on frequency. Broadband phase measurements at the feed terminals can be referenced to a known phase center at a particula frequency, and used to predict variations in the feed radiation pattern with frequency.

From known feed amplitude, phase, and polarization data, antenna pattern parameters for a single frequency may be calculated using known computational procedures [5, 15]. For a paraboloidal reflector, the amplitude of the aperture field, $\underline{E}_a$, is related to the amplitude of the feed pattern $\underline{E}_f$ by

$$|\underline{E}_a(\theta,\varphi)| = \left(\frac{1 + \cos\theta}{2}\right) |\underline{E}_f(\theta,\varphi)| \tag{53}$$

where θ and φ are conventional coordinates as illustrated by Figure 29. The aperture phase function $\Phi(\rho,\varphi)$ is determined from the phase of the feed patterr by

$$\Phi(\rho,\varphi) = \tan^{-1}\left[\frac{\mathrm{Im}\ \underline{E}_f(\theta,\varphi)}{\mathrm{Re}\ \underline{E}_f(\theta,\varphi)}\right], \tag{54}$$

where $\rho = \dfrac{2L \sin\theta}{1 + \cos\theta}$. Recently, a method for determining the aperture polarization by a sterographic mapping of the feed polarization has been developed [6]. Thus, techniques for determining aperture amplitude, phase, and polarization from single frequency feed data have been developed.

Once the complex aperture distributions $\underline{E}_a$ and $\underline{H}_a$ have been determined, the far-zone fields can be obtained by several methods. These methods generally differ in their computational complexity and in the completeness of data obtainable. The most accurate far-field data are obtained by using a vector formulation [7]. However, the evaluation of the required vector integrals is tedious and very time consuming for electrically large apertures. A less accurate but more tractable approach may be to calculate the far-zone using an approximate scalar relation. The scalar relation is in the form of a two-dimensional Fourier integral which can be efficiently evaluated using

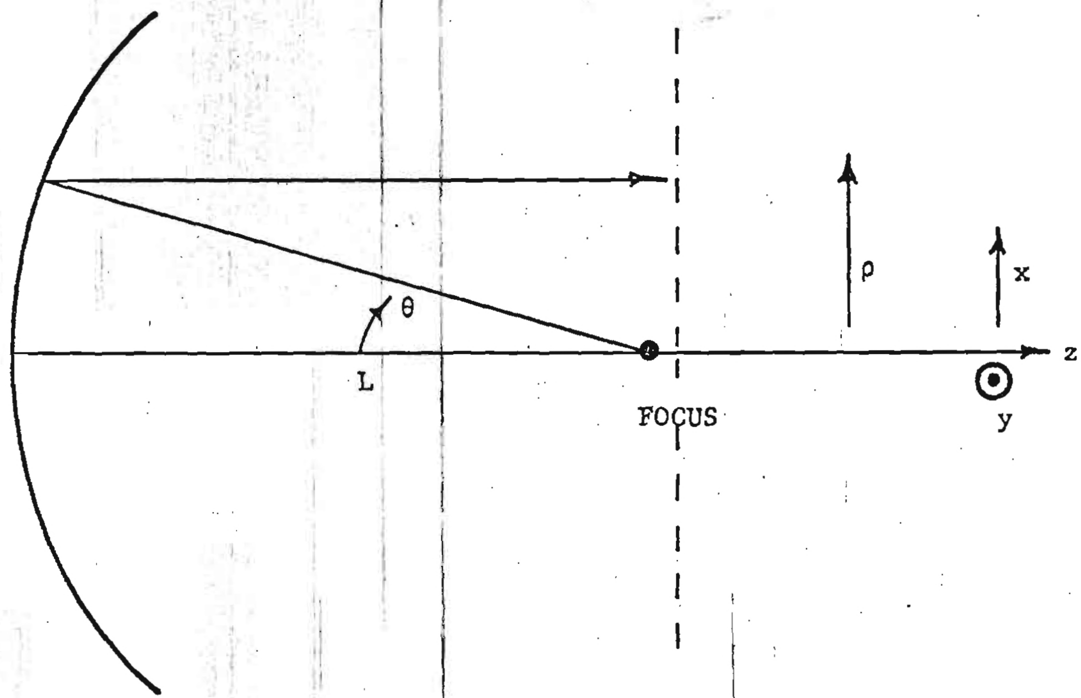


Figure 29. Paraboloid of revolution geometry

the Fast Fourier Transform (FFT) algorithm. In order that the scalar evaluation be accurate, there are several restrictions which must be met [7]. The effect of these restrictions in the present application must be quantified.

Another method which may prove useful for calculation of far-removed sidelobes employs the high-frequency asymptotic methods of the geometrical theory of diffraction (GTD) [8, 9, 10]. Because the GTD method is presently not as accurate in the main beam region as in the sidelobe regions, the previously discussed integral formulations must be employed in the main beam region. Further, because previous applications of these computational procedures have been limited to narrow-band antennas, the feasibility of applying these procedures to wideband antennas must be investigated.

It may be seen that there are several problem areas to be investigated in order to define the feasibility of the three-step approach discussed above. The most efficient and direct method of obtaining feed radiation parameters from broadband phase and amplitude data must be determined. The feasibility of predicting broadband far-field parameters with methods previously used in single frequency analysis must be investigated. The required computational complexity (computer resources) and the resulting accuracy of the various methods must be determined.

SECTION V

CONCLUSIONS AND RECOMMENDATIONS

The objectives of this contract were to explore new techniques to measure, record, and analyze broadband antenna patterns and to determine the feasibility of developing instrumentation to perform these measurements. These objectives have been fully met, and the investigations which have been described in this report establish the feasibility of broadband phase and amplitude measurement systems which are applicable to high-gain microwave antennas. Specific system concepts which were to be investigated included amplitude-only broadband noise systems and sweep frequency phase and/or amplitude systems. Variations from these two basic concepts were also to be investigated. Evaluations showed that the broadband noise concept is not a cost-effective approach for an amplitude-only system. A simpler sweep frequency system (called a Hybrid System) which employs electronic integration of a sweep frequency signal to obtain broadband amplitude data was synthesized and evaluated. This Hybrid system is an effective approach to an amplitude-only system. Preliminary design of the amplitude-only Hybrid system has been completed, and its construction is relatively straightforward with available components.

A computer-controlled sweep frequency phase-plus-amplitude system is required to completely characterize an antenna that requires a wide instantaneous bandwidth and to provide the necessary data to perform off-line diagnostics which define that antenna's frequency and time-domain performance under a wide variety of operating parameters. System operating parameters (frequency and spatial sample rates) and calibration techniques which permit a high measurement accuracy are practical. Although the computer controlled sweep frequency phase/amplitude system is relatively complex, the individual components required for its instrumentation are feasible with today's technology. The primary hardware advance which is required involves incorporating a broadband phase-locked frequency tracking capability into an existing wide-range receiver.

Because of the relative simplicity of the Hybrid system, it can be constructed with a very modest expenditure of time and money. Construction of this amplitude-only system should begin immediately, thus demonstrating at an early date a significant increase in broadband antenna measurement capability. Amplitude-only spatial antenna patterns impact on a number of radar performance parameters (detection range, angular resolution, tracking accuracy, ECM susceptibility) and the ability to quickly measure these patterns over a broadband is greatly needed.

Because there are important applications in which amplitude-only information is not sufficient (such as off-line diagnostics to predict pulse distortion, and effects on FM modulated signals), completion of the tasks which are necessary to develop the computer-controlled sweep frequency phase-plus-amplitude system should be pursued. These additional tasks include (1) complete system design and component specification, (2) complete definition of system operating procedures, with specification of procedures for any required computer correction of both amplitude and phase data, (3) develop software necessary for computer-control and data recording with an operator-interactive capability, (4) breadboard a laboratory setup and experimentally demonstrate feasibility of the system, (5) analytically bound the reference antenna calibration accuracy which is practical, and (6) develop and demonstrate computer software to compute an antenna's effect on an arbitrary input pulse from measured phase/amplitude data on the antenna.

Because of prohibitive antenna size or other restraints on moving an antenna, in some important cases it may not be practical to measure antenna pattern parameters directly. Consequently, in addition to the above basic tasks, the feasibility of predicting broadband far-field parameters of electrically large apertures from measured feed phase/amplitude data should be investigated. Besides the antenna measurements problems of immediate interest to RADC, there are many other applications of broadband phase and/or amplitude measuring systems. Some of these other applications are identified below, along with the respective cognizant agency.

APPLICATION	AGENCY
Spread Spectrum Radar	U. S. Army MICOM Huntsville, Alabama
Amplitude and Phase Characterization of Up to 10:1 Bandwidth Antennas	U. S. Army, MICOM Huntsville, Alabama
RATSCAT Radar (330 pico-sec pulsewidth)	U. S. Air Force Holloman AFB, New Mexico
Test Wideband Antennas for Guidance System	Naval Weapons Laboratory Dahlgren, Virginia
Mine Detection Radar	U. S. Army MERDC Fort Belvoir, Virginia
Short Pulse Transmitters	RADC Griffiss AFB, New York
Real Time Pulse Compression	RADC Griffiss AFB, New York

It may be concluded that there is great need for broadband phase and/or amplitude measuring systems. To fill this need, the following specific actions are recommended:

(1) construct a complete octave bandwidth (2-4 GHz) Hybrid system and verify its operation through laboratory test,

(2) complete system and component specification of the computer-controlled sweep frequency phase/amplitude system,

(3) completely develop system operation and computer correction procedures for the phase/amplitude system, including computer correction for reference antenna calibration,

(4) develop necessary operating software for the phase/amplitude system,

(5) demonstrate system feasibility through laboratory test,

(6) develop and demonstrate with selected examples the software required to predict an antenna's effect on an arbitrary input pulse, and

(7) investigate the feasibility of predicting broadband far-field parameters of electrically large apertures from measured antenna feed data.

These recommended actions will lead to a realization of the complete phase/amplitude system in an efficient manner with an early demonstration of a broadband amplitude-only system with a significant broadband measurement capability.

SECTION VI

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SECTION VII

APPENDICES

APPENDIX I

COMPONENT DESCRIPTIONS

A. CONCEPT I SYSTEM COMPONENTS

1. General

The major components peculiar to implementation of a broadband noise system are discussed here. Other required components for actual assembly of a system are discussed in Section II, Concept I Systems. Since all types of broadband systems which have been investigated require a broadband transmitting antenna, antennas are discussed separately in Subsection C of this appendix.

2. Noise Sources

The noise power P_n available from a thermal noise source is given as

$$P_n = kTB, \tag{A-1}$$

where

k = Boltzmann's constant = 1.38×10^{-23} joules/degree Kelvin,

T = temperature in degrees Kelvin, and

B = bandwidth in Hertz.

The noise power density output of a resistor at room temperature (290°K) is therefore - 114 dBm/MHz.

Power output of commercially available noise sources is specified by the term Excess Noise Ratio (ENR), defined as

$$ENR(dB) = 10 \log \frac{k(T - T_0)\ B}{k\ T_0\ B}, \tag{A-2}$$

where

kTB = available noise power at operating temperature,

kT_oB = available noise power at 290°K, and

T = equivalent absolute temperature of the noise source.

In order for the noise source to be useful, T must be greater than twice T_o. Generally T is much greater than T_o. Thus, the noise power output of a commercial noise source is determined from the following equation:

$$P_n = -114 + ENR + 10 \log (\text{Bandwidth in MHz}) = \text{noise power in dBm.} \quad \text{(A-3)}$$

Typical data for various types of noise sources are shown in Table 7. For an ENR of 15 to 30 dB and a bandwidth of 1 GHz, corresponding noise power output would be from -69 to -54 dBm.

TABLE 7. TYPICAL NOISE SOURCES

TYPE	EXCESS NOISE RATIO (dB)
Temperature - limited Thermionic diode	6
Hot Wire Resistor (upper limit)	11
Argon discharge tube	15
Semiconductor diode	30

Thermal noise sources, by their inherent physical nature, generate noise power which has a uniform spectral distribution (white noise) over a frequency range well in excess of the frequencies of interest here. Bandwidth limitation is due to the transmission lines necessary for coupling. Coaxial cable coupled noise sources have uniform spectral distribution from VHF to about 12 GHz, while waveguide coupled ones are limited by the bandwidth of the particular waveguide used. The upper frequency limit of readily available noise sources appears to be about 40 GHz.

Since the transmitter power requirement here is on the order of 10 watts (+40 dBm), with -69 to -54 dBm of noise power available a gain of from 94 to

109 dB is necessary to amplify the noise source output to the required power level even if all other system losses are neglected. Amplifiers for this function are discussed in the following subsection.

3. Traveling Wave Tube (TWT) Amplifiers

As indicated above, amplification of from 94 to 109 dB is required to bring the noise source output to the proper power level. Both tube and solid-state microwave amplifiers are available. However, at the present state-of-the-art, obtaining 10 watts output with octave band coverage is not feasible with solid-state microwave amplifiers above about 1 GHz. Of the tube amplifiers, TWTs are well suited to this application since they have octave bandwidths along with high gain and more-than-adequate power capability.

Although each major element of a TWT performs a simple function, several different implementations of these elements are possible, and consequently, tubes of differing characteristics result. The major elements of a TWT are (1) the electron gun, (2) the RF interaction circuit, (3) the electron beam focusing device, and (4) the collector. Major differences in common TWTs are accounted for primarily by the choice of RF interaction circuit and beam focusing device. For broadband TWTs, the helix RF circuit is most appropriate, and bandwidths of 5:1 have been obtained. Beam focusing is accomplished by means of either permanent magnets or solenoid electro-magnets. Permanent magnet focusing is most convenient since no additional power supply is required. Electro-magnets are generally used in very high power tubes, but permanent magnet tubes have more than adequate power capability (up to several kw) for the present application. Thus, the most suitable power amplifier for this application is a TWT that utilizes a helix interaction circuit and permanent magnet focusing. Tube parameters of interest include gain, bandwidth, power output, and dynamic range. Typically available octave bandwidth TWTs with a 10-watt output capability have gains from 35 to 50 dB, depending on frequency. Thus, because 94 to 109 dB of amplification is required, two or three stages of gain might be required to obtain 10 watts of noise output. In addition to amplifying the input signal, TWTs also generate excess noise. In particular, a high power TWT generates considerable noise. However, the noise available from a TWT does not have a broadband uniform spectrum. Hence, to preserve the required noise properties throughout the transmitter chain, the uniform noise spectrum input at each TWT stage should be large relative to the excess noise generated by the TWT. A low-noise TWT must be used in the low level portion of the transmitter.

It is within the state-of-the-art to produce TWTs at the required power level and with octave or waveguide bandwidths up to beyond 75 GHz. However, between about 18 GHz and 75 GHz, due to lack of demand, tubes may not be readily available for certain frequency ranges.

4. Broadband Receivers

The bandwidth of the receiver must be as wide as that of the transmitter in order to accurately measure the response of the test antenna. One of the properties of broadband levelled noise is that over its frequency band, it has constant spectral density, i.e., it has constant power per unit bandwidth. If this signal were passed through a frequency selective device such as the test antenna, the spectrum at the output of the device would correspond to the frequency response characteristics of that device. When this spectrum is in turn detected by a broadband square-law detector, the voltage produced is proportional to the total RF power within the spectrum bandwidth. Thus, frequency integration of the response characteristics of the device occurs and is expressed as

$$V_o = \Phi K_d \int_{S_1}^{S_2} H(S)dS \quad , \tag{A-4}$$

where,

Φ = constant amplitude incident power spectrum,

$H(S)$ = amplitude transfer function of the device (function of frequency)

dS = incremental frequency bandwidth,

K_d = detector constant, and

V_o = voltage output from the detector.

Thus, the detector output is proportional to the integral of the device frequency response characteristics. Therefore, to accurately measure antenna response over the full transmitted spectrum bandwidth, a receiver of equal bandwidth is required.

General types of wideband receivers include the following:

(a) detector-video receiver,

(b) wideband RF receiver,

(c) heterodyne receiver with CW local oscillator and wideband IF, and

(d) heterodyne receiver with swept local oscillator and narrow-band IF.

Each of these receiver types will be discussed individually.

A detector-video receiver is one in which the incoming RF signal is converted directly to video by a detector, and the resulting video signal is amplified by a high gain video amplifier. The RF detector is usually a square-law device, i.e., a device that produces an output voltage proportional to the input RF power. Either a semiconductor diode or a thermal device such as a barretter may be used as the square-law detector. Semiconductor diodes have about 20 dB more sensitivity than the barretter, but the diodes are generally limited to about 40 dB in dynamic range. However, post-detection techniques can be used to extend the dynamic range to 60 or 70 dB by processing the video signal. This type of receiver is extremely simple, the RF bandwidth is limited only by the detector, and the entire frequency range of interest (1-15 GHz) can be covered by a single detector. However, it has the lowest sensitivity of the four types.

The wideband RF receiver consists essentially of the above described detector-video receiver preceded by a low-noise wideband RF amplifier. This type of receiver has a very high sensitivity, but typically a separate amplifier is required for each octave or waveguide band of interest. Several different types of low-noise RF amplifiers are available. These include low-noise TWTs, transistor amplifiers, tunnel diode amplifiers, and parametric amplifiers. At the lower frequency ranges (up to X-band), the transistor amplifier offers good noise performance and broad bandwidth at a relatively low cost. Single transistors covering the frequency range from 200 MHz to 1 GHz are available, but over the frequency range of 1 GHz to 8 GHz, octave bandwidths are typical. Low-noise TWTs also offer octave bandwidths, but their noise figures at frequencies below X-band are generally slightly higher than those of the transistor amplifiers. At X-band or above, an octave bandwidth is difficult to achieve with any of the commonly available low-noise amplifiers except the TWTs. Above the frequency range of about 18 to 20 GHz, low-noise TWTs generally cover standard waveguide bands, up to 75 GHz. Thus, the best approach is a low-noise transistor amplifier up to 8 GHz with a low-noise TWT at 8 GHz and above.

In a conventional heterodyne receiver, the input signal is converted to an intermediate frequency with the same bandwidth regardless of the RF frequency. However, down-conversion is impractical for this application because an equal frequency range at a lower center frequency results in increased fractional bandwidth which would require an impossibly broad IF amplifier for the bandwidth under consideration here. Up-conversion would result in a smaller percentage IF bandwidth, but because the IF amplifier would be at a very high frequency, it also is not a good solution.

A heterodyne receiver with swept local oscillator and narrow-band IF amplifier is incompatible with Concept I because the instantaneous receiver bandwidth would not be great enough to process the broadband noise signal without loss of information. However, this type of receiver can be made compatible with Concept I by the use of additional hardware to process the detected video signal (a synchronized integrator and a sample and hold circuit)

Thus, the most straightforward receiver design for the broadband noise transmitter is a wideband RF receiver with a separate amplifier for each frequency band of interest.

5. Pattern Recorder

The selected broadband receiver must be compatible with the Scientific Atlanta, Inc., Series 1520 Rectangular Coordinate Pattern Recorder. This recorder is used in antenna testing to plot the received signal as a function of the angle of the test antenna. When this recorder is fed with a square-law detector (as described for the wideband RF receiver), either received power, received voltage, or their logarithms may be plotted. With a square-law detector, the recorder has a 40-dB dynamic range. When the recorder is to be used with a detector receiving an audio modulated carrier (e.g., 1 kHz), it is equipped with a narrow-band, crystal-bolometer amplifier. For recorder operation from a dc input signal, the crystal-bolometer amplifier is replaced with a dc-chopper preamplifier. Thus, the 1520 recorder can interface with practically any type of receiving system when furnished the appropriate amplifier and drive level.

B. CONCEPT II SYSTEM COMPONENTS

Concept II systems, which rely on phase and/or amplitude samples over a wide frequency range, require a basic signal source which can be rapidly tuned over the desired measurement range. In addition to this basic signal source, the other major component of the transmitter is a microwave power amplifier. Major components required at the receiving site are a wideband phase and amplitude receiver, a computer for system control and interface, and data recorders. These major Concept II system components will be discussed in this section. Broadband transmitting antennas are discussed in Subsection C.

1. Signal Sources

Signal sources appropriate for Concept II systems include both sweep oscillators and frequency synthesizers. Sweep oscillators are more versatile for test instrumentation, and they provide greater frequency coverage for a given expenditure. However, the frequency synthesizer generally provides greater frequency stability, less phase noise, and consequently greater phase measurement accuracy with the group delay type of system. Hence, a frequency synthesizer may be required in the phase/amplitude system. Both types of signal sources are described below.

The microwave frequency synthesizer typically uses a YIG-tuned oscillator, and has its output derived directly from this source. YIG tuning is employed both for signal purity and linear tuning. Octave or greater tuning ranges are typical with linearity on the order of 0.1% and spurious radiation depressed by at least 60 dB below the desired signal. Other important features of frequency synthesizers are their frequency stability and freedom from phase noise. Phase noise (in a 1 Hz bandwidth 100 kHz from carrier) is typically 90 dB below the carrier, and frequency stability is on the order of 1×10^{-8}/day. Frequency resolution of less than 100 kHz is standard. Octave bandwidth units from 1 GHz to 8 GHz are presently available with coverage to about 18 GHz practical at the present state-of-the-art.

Sweep oscillators, by eliminating the need of tedious point-by-point testing, enable practical broadband evaluation of frequency-dependent parameters, i.e., amplitude and phase. Many modern sweep oscillators use solid-state devices as the power source and use a modular construction in which the RF source is a plug-in unit. Plug-ins which cover the ranges from a few MHz to over 1 GHz, octave bands from 1 to 8 GHz, and waveguide bands above 8 GHz have been available for several years. However, systems are now available in which the outputs from individual sources are electronically multiplexed so that coverage of the entire 1-18 GHz range is provided at one output. Standard features of commercially available sweep oscillators include internal or external leveling, variable sweep widths, internal or external modulation capability, selectable sweep mode (single sweep, repetitive, triggered, or manual), single-frequency CW operation, variable sweep times (10 msec per sweep to 100 sec per sweep), and remote programability. Above 18 GHz, signal sources covering standard frequency bands (18-26.5 GHz, 26.5-40 GHz, 33-50 GHz, and 50-75 GHz) are catalog items. However, due to less demand for test instruments in these frequency ranges, these devices have not been designed and manufactured into such versatile sweep oscillator sources as described for the lower frequency bands. Moreover, these higher frequency sources are less readily available, and they are generally more expensive than the lower frequency ones. In some cases, it would be practical to use a versatile lower frequency sweep oscillator followed by a frequency multiplier to obtain the desired test signal at a higher frequency.

2. Broadband Amplifiers

Since the signal sources described above typically have power outputs on the order of 1 mW, amplification of 40 dB is required to obtain the desired 10 watts of transmitter power. Up to 1 GHz, wideband transistor amplifiers with a 10-watt power output capability are available. For example, over the frequency ranges of 50 MHz to 500 MHz and 500 MHz to 1 GHz, 40-dB gain, 10-watt units are available. Above 1 GHz, transistor amplifiers cannot presently provide the required power and bandwidth combination. However, as discussed in Subsection A-3, TWT amplifiers of octave bandwidths whose power capabilities and gains are more than adequate are readily available up to about 18 GHz. Between 18 and 75 GHz, tubes offering waveguide band coverage have been demonstrated, although tubes covering some particular frequency ranges may not be readily available.

The phase stability of the amplifiers, in addition to the power, bandwidth, and gain, will be important for Concept II systems. Transistor amplifiers have demonstrated phase linearity and stability in phased-array applications (including pulse compression). In TWTs, phase shift is sensitive to such parameters as beam voltage, drive level, and heater voltage. Therefore, achieving a stable phase shift in the TWT requires regulation of these parameters. The most critical factor is that of beam voltage. A typical TWT exhibits a 3° phase shift for a 0.1% change in beam voltage. Regulation to this level or better is available in standard amplifier units suitable for this application. Consequently, suitable signal sources and amplifiers for the 200 MHz to 75 GHz range are feasible.

3. Receivers

Two different classes of commercially available receivers should be considered for sweep-frequency phase/amplitude systems. These classes are

(1) superheterodyne phase-locked receivers and (2) amplitude and phase network analyzers. These receivers were designed for different types of measurement problems, and therefore, they have differing characteristics. However, they have some overlap in their applications, and the feasibility of using either a standard or a modified version of both of these receiver classes was considered. (Amplitude-only spectrum analyzers were also considered initially, but it was concluded that a sweep-frequency computer-controlled amplitude-only system was not cost-effective and that spectrum analyzers are not applicable.)

Network analyzers and microwave receivers are both capable of providing the output phase and amplitude of a signal in a visual display and digital formats. There are several significant design differences between typical receivers and network analyzers (such as data bandwidths, methods of obtaining the LO signals, and detection methods), but the primary operational difference between the two types of devices is their sensitivity. Depending on the frequency range and bandwidth of interest, the sensitivity of network analyzers typically is 10 to 40 dB poorer than the sensitivity of microwave receivers. Network analyzers have been used for short-range antenna measurements (for example, in an anechoic chamber) by locating the reference signal generator and the network analyzer at the transmitting site and then routing the received test signal back to the network analyzer via a coaxial cable. Because of the requirement that the test and reference inputs must be fixed relatively close together, coupled with its poorer sensitivity, the network analyzer is not suited to the applications considered here. Although no commercially available microwave receiver currently has the desired swept frequency phase and amplitude capability, the modification of an existing microwave receiver to provide the required measurement capabilities is feasible. Scientific-

Atlanta, Inc., has defined the changes necessary to modify the Scientific-Atlanta Series 1750 Broadband Phase/Amplitude Receiver and to provide a tracking sweep-frequency measurement capability. In addition, the modifications required to also provide a group delay measurement capability have been investigated. These design modifications and the resulting receiver specifications are fully described in Section IV.

4 Minicomputer/Tape Recorders

The Concept II systems include a small computer as a data acquisition and storage and process control device. Such small computers are commonly termed minicomputers. Table 8 lists some typical characteristics of presently available minicomputers. Since the hardware, software, and architecture of these machines are in a present state of rapid commercial development, only general specifications are listed and these are often rapidly outdated. All of the features in Table 8 are available either as standard features or as standard options on the typical minicomputer.

The first five features of Table 8 are critical minicomputer requirements for real-time antenna measurements.

TABLE 8. COMMON MINICOMPUTER FEATURES

- Program Interrupt
- Real-Time Clock
- Direct Memory Access (DMA)
- 4K-64K Memory Incorporating:
  - Fast Access Core Storage
  - Read-Only Memory (ROM)
  - Programmable ROM
  - Internal Registers
  - Scratch Pad Memory
- Versatile Input/Output (I/O) Interconnection
- Cycle Time: 400 ns - 2.5 μs
- Word Length: 12-18 Bit
- Direct and Indexed Addressing of Storage
- "Hardwired" Arithmetic Units

The program interrupt feature allows Input/Output (I/O) devices to temporarily suspend an operating program so that I/O can be processed as required. The function of the real-time or external clock is to provide interrupts at prescribed rates. Thus, data can be acquired at precise sampling rates, or the CPU can execute a predetermined action at the correct instant in time. The direct memory access feature allows data to be transferred into and out of memory directly by allowing the I/O device to delay the CPU processing by one machine cycle while entering a single data word. This feature increases the data transfer rate. This can be significant when a large amount of data is required. The minicomputer should also provide for the interconnection of a variety of Input/Output devices. A specific I/O device required for the antenna measurement system is a digital tape unit for storage of measured data for subsequent processing. All minicomputers conform to the industry standard tape format. Tapes recorded on these machines interface with all large-scale computers which conform to the industry standard. Consequently, the Raytheon 703 Minicomputer for on-line operation and the Honeywell 600-line computers for off-line processing are both compatible with the Hewlett-Packard 2020E Digital Magnetic Tape Recorder.

C. TRANSMITTING ANTENNAS

A broadband transmitting antenna with selectable polarization (linear vertical, linear horizontal, or circular polarization) is desirable with all the systems which were considered in this study. In addition, the systems which have been selected for further development may require a reference antenna at the receiving site. With the Hybrid System, compensation for variations in power density at the receiving aperture is based on use of a reference antenna, and both of the phase/amplitude receivers require a frequency tracking channel. There are three general types of microwave antennas for broadband operation: (1) frequency independent antennas such as the spiral and log periodic, (2) broadband horn antennas, such as the ridged horn or pin-wall, and (3) reflector antennas with broadband feeds. To obtain the desired gain, a reflector antenna with a broadband feed is the most appropriate choice for the current applications. Reflector antenna systems with good performance properties over octave or greater bandwidths can be achieved. Due to the increase of its electrical size, the gain of a reflector antenna increases by 6 dB per octave of frequency increase, but the feed horn effects may modify this change somewhat. However, since the space loss also increases by the same factor, the two effects tend to cancel. Consequently, this method would produce a more nearly constant power density at the receiving antenna than that which would be produced if a transmitting antenna with constant gain versus frequency were used. Because both the Hybrid and the phase/amplitude systems can accommodate changes in power density, it is not critical that the two effects do not exactly cancel over an octave bandwidth. The Hybrid system electronically compensates for these gain changes, while with the phase/amplitude system, calibration data will be used to computer-correct the measured test data and account for any variations in power density at the test antenna.

To limit ground reflections, a narrow-beam transmitting antenna is desired. Further, a fairly high gain transmitting antenna is desirable to produce an adequate received power level without an excessively high power transmitter. Calculations show (see Section III) that detection requirements can be met if the reflector is sized to provide a 30-dB gain transmitting antenna at 4 GHz. If the reflector is sized such as to provide 30 dB of gain at 4 GHz, the change of gain with frequency will compensate for the change of space loss with frequency. At frequencies below 4 GHz, the space loss decrease to compensate for the reduction in gain; however, as the frequency is reduced from 4 GHz, the beamwidth of the reflector antenna broadens. Hence, for a reflector that has been sized at 4 GHz, the possibility of ground reflections increases as the frequency is reduced. Selectable polarization can be provided by employing a selectable polarization feed with a paraboloidal reflector. Common feed types for obtaining dual polarization include log-periodic and quad-ridge horns. These feeds, which can typically provide either dual linear (vertical and horizontal) or dual circular (right and left handed) polarizations, have good performance over octave or greater bandwidths. External power dividing and switching can be used with these dual polarized feeds to provide two orthogonal linear and two orthogonal circular polarizations from a single feed. Perhaps a more practical approach is to provide both a dual linear feed/reflector system and a dual circular/reflector system and then switch between the systems as required. Greater bandwidth and increased polarization isolation can be expected from the latter approach. Data on representative dual polarized feeds are shown in Table 9.

TABLE 9. BROADBAND DUAL POLARIZED FEED DATA

Feed Type	Frequency Range GHz	Polarization
Quad-Ridge Horn	2-4	Dual linear
Quad-Ridge Horn	2-8	Dual linear
Quad-Ridge Horn	4-8	Dual linear
Quad-Ridge Horn	8-16	Dual linear
Log-Periodic	12-18	Dual linear
Log-Periodic	1-12	Circular
Log-Periodic	12-18	Circular
Log-Periodic	0.3-1	Dual linear
Log-Periodic	1-4	Dual linear

APPENDIX II

TEST FOR IDENTICAL ANTENNAS

This appendix presents the mathematical relations which allow one to define conditions for determining if two antennas are electrically identical. The fundamental equations which describe the performance of an antenna as a transmitter, receiver, and scatterer are presented in Appendix III. The circuit analog of an antenna's transmitting and receiving properties can be represented by a Thevenin equivalent circuit. However, the scattering properties cannot be described by a simple Thevenin equivalent since it is not generally possible to compute the scattered power from such a Thevenin equivalent circuit. The equivalent circuit of Figure 30 can be used as a circuit analog for antenna scattering [11].

In this circuit, V is the open circuit Thevenin equivalent voltage, Z_a is the antenna impedance, Z_ℓ is the load impedance, Z_1 and Z_2 are two impedances to be determined such that the voltage V_s between terminals 1 and 2 is proportional to the scattered field, and V_ℓ is the load voltage. The terminal voltage V_s may be expressed as

$$V_s = - \frac{V Z_2}{Z_1 + Z_2} + \frac{Z_\ell V}{Z_\ell + Z_a} \quad . \tag{A-5}$$

Since the load current I_ℓ is given by

$$I_\ell = - \frac{V}{Z_a + Z_\ell} \quad , \tag{A-6}$$

and the scalar form of the scattering equation is

$$E^s{}_\ell (Z_\ell) = E^s(0) - \frac{Z_\ell \, I(0)}{Z_\ell + Z_\ell} E^r \quad , \tag{A-7}$$

where

$$E^s(0) = E^s{}_\ell(Z_\ell) \ @ \ Z_\ell = 0, \text{ and}$$

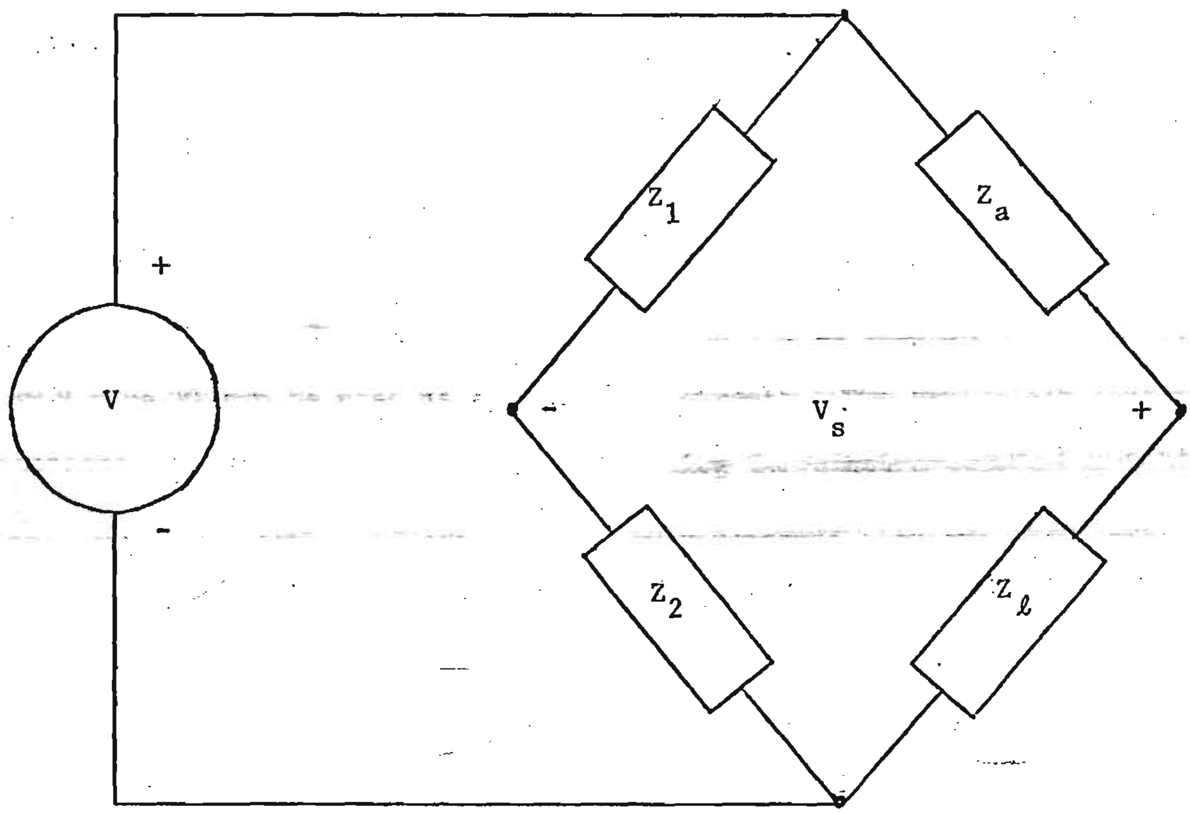

Figure 30. Equivalent bridge circuit for antenna scattering

$$I(0) = I_\ell @ Z_\ell = 0,$$

one has

$$Z_a \frac{E^s(Z_\ell)}{E^r} = \frac{Z_a E^s\,(0)}{E^r} + \frac{Z_\ell V}{Z_\ell + Z_a} \quad . \tag{A-8}$$

If one lets

$$\frac{Z_a\, E^s(0)}{E^r} = - \frac{V\, Z_2}{Z_1 + Z_2} \quad , \tag{A-9}$$

then

$$E^s(Z_\ell) = \frac{E^r}{Z_a} V_s \quad , \tag{A-10}$$

where E^r is the radiated field strength per unit current. From Equations A-6 and A-7 one has

$$\frac{E^r}{Z_a} = - \frac{E^s(0) - E^s\,(\infty)}{V} \quad , \tag{A-11}$$

where

$$E^s(\infty) = E_\ell{}^s(Z_\ell) @ Z_\ell = \infty \; .$$

From Equations A-11 and A-9, one gets

$$\frac{Z_2}{Z_1} = - \frac{E^s(0)}{E^s(\infty)} \quad . \tag{A-12}$$

This ratio relates the scattered field to the open circuit Thevenin equivalent voltage by the circuit of Figure 30 and Equation A-10. The relation given here can be used to determine whether two antennas are electrically identical. With respect to transmission and reception, two antennas whose impedance Z_a is identical as a function of frequency are identical. With respect to scattering, Equation A-12 must be satisfied by both antennas. Thus two antennas are electrically identical at a given angular frequency ω

and aspect angle if

1) their input impedances are equal, and

2) the ratio of the scattered field for short and open circuit load conditions is equal.

A measurement of $Z_a(\omega)$ and $\dfrac{E^s(Z_\ell = 0,\ \omega)}{E^s(Z_\ell = \infty,\ \omega)}$ for all aspect angles of interest is required to determine whether two physically similar antennas are electrically identical.

APPENDIX III

THE ANTENNA AS A TRANSMITTER AND RECEIVER

The electric field $\bar{E}$ at a great distance from an antenna is transverse to the direction of propagation and can be written as

$$\bar{E} = \hat{\theta}E_{\theta} + \hat{\varphi}E_{\varphi} = -j \frac{\eta I_t}{2\lambda} \bar{h}_t \frac{e^{-jkr}}{r} , \tag{A-13}$$

where

η = impedance of free space = 376.7 ohms,

I_t = transmitting antenna input-terminal current,

λ = wavelength,

$\bar{h}_t$ = effective transmitting antenna height,

$k = \frac{2\pi}{\lambda}$,

and r, θ, φ are conventional spherical coordinates. The effective transmitting antenna height contains both the polarization and pattern information. For the transmission of EM energy between a pair of antennas, the electric field $\bar{E}^i$ at the receiving antenna, at a range r from the transmitter, is given by Equation A-13, and this field induces an open circuit voltage at the receiver terminals given by [12]

$$V_r = -\bar{h}_r \cdot \bar{E}^i ; \tag{A-14}$$

where $\bar{h}_r$ is the effective receiving antenna height.

The circuit analogy of the receiving properties of the receiving antenna can be represented by the Thevenin equivalent circuit shown in Figure 31. The current I_r through the load Z_{ℓ} on the receiver terminals can be obtained from a Thevenin circuit as

$$I_r = - \frac{V_o}{Z_a + Z_{\ell}} . \tag{A-15}$$

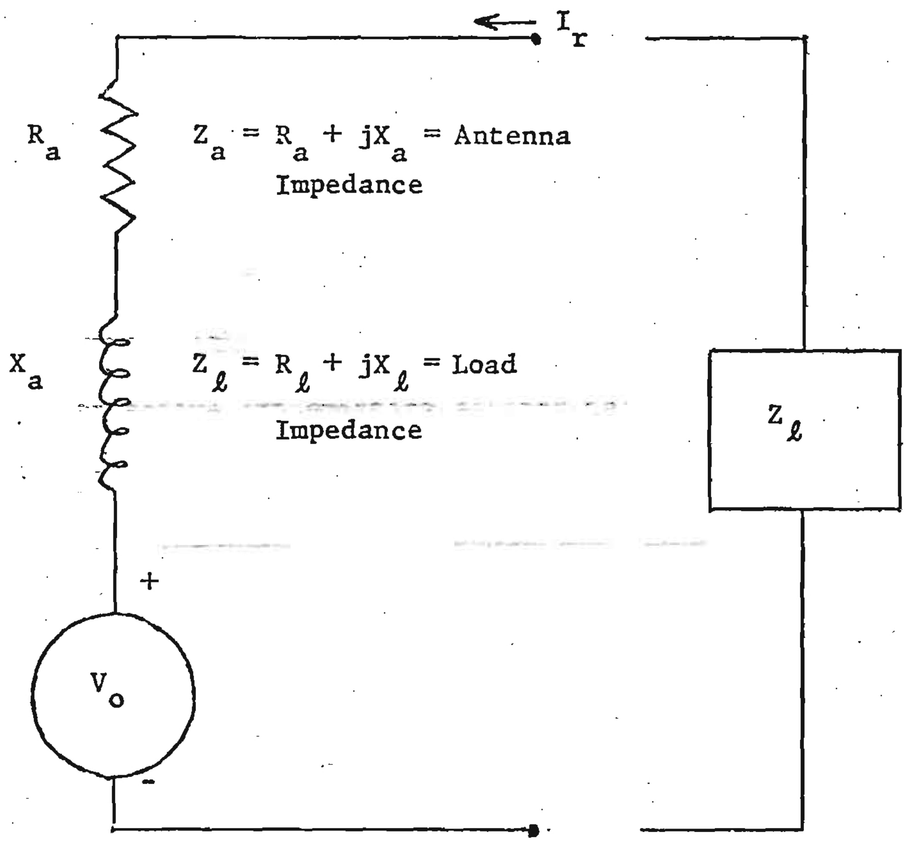

Figure 31. Thevenin equivalent circuit antenna analog

Thus the voltage V_r across the load Z_ℓ is given by

$$V_r = (j \frac{\pi}{2\lambda}) \frac{e^{-jkr}}{r} I_t (-\bar{h}_r \cdot \bar{h}_t) \left(\frac{Z_\ell}{Z_a + Z_\ell}\right) . \tag{A-16}$$

The transmitting antenna terminal current I_t is related to the voltage V_t impressed on the terminals by the transmitting antenna impedance Z_t by

$$V_t = I_t Z_t . \tag{A-17}$$

Thus the ratio between the transmitted and received voltage is given as

$$\frac{V_r}{V_t} = (j \frac{\pi}{2\lambda}) \frac{e^{-jkr}}{r} (-\bar{h}_r \cdot \bar{h}_t) \left(\frac{Z_\ell}{Z_a + Z_\ell}\right) \frac{1}{Z_t} = S_{12} . \tag{A-18}$$

We can relate this expression to the usual definition of the antenna gain by using the Friis transmission formula,

$$\frac{W_r}{W_t} = \frac{A_{er} A_{et}}{\lambda^2 r^2} = \left[\frac{\lambda^2}{4\pi} D_r\right] \left[\frac{\lambda^2}{4\pi} D_t\right] \frac{1}{\lambda^2 r^2} , \tag{A-19}$$

where A_{er}, A_{et} = effective aperture of the receiver and transmitter, respectively,

D_r, D_t = directivity of the receiver and transmitter, respectively, and

W_r, W_t = power received and power transmitted, respectively.

Assuming a lossless antenna structure, the power gains and the directivities are equal so that $G_r = D_r$ and $G_t = D_t$, with the result that

$$\frac{W_r}{W_t} = \left(\frac{\lambda}{4\pi r}\right)^2 G_r G_t = \left|\frac{E_r}{E_t}\right|^2 . \tag{A-20}$$

Thus, the antenna voltage gain transfer is given by

$$\left|\frac{V_r}{V_t}\right| = \frac{\lambda}{4\pi r} \left|G^v{}_r\right| \left|G^v{}_t\right| . \tag{A-21}$$

If the two antennas are identical, $Z_a = Z_t$, and if they are aligned such that $\bar{h} = \bar{h}_t (\theta, \varphi) = \bar{h}_r (\theta, \varphi)$, i.e., no polarization or pattern mismatch, one has from Equation A-18

$$S_{12} = \left(j \frac{\eta}{2\lambda} \frac{e^{-jkr}}{r} \right) (h)^2 \frac{Z_\ell}{Z_a + Z_\ell} \cdot \frac{1}{Z_a} \quad . \tag{A-22}$$

The scattering parameter S_{12} can be measured using a network analyzer scatterin parameter test set. If the range r is known $h^2(\omega)$ and thus $h(\omega)$ can be determined. Once $h(\omega)$ is known, the phase and amplitude properties of the antenna can be calculated. The transmitted field due to an excitation current I_t or the received voltage due to an incident electric field $\bar{E}^i$ can be determined using Equations A-13 and A-14, respectively.

APPENDIX IV

THE ANTENNA AS A SCATTERER

This appendix presents the fundamental equations which describe an antenna as a free space scatterer. The Thevenin equivalent circuit which represents the antenna when acting as a receiver is not sufficient to fully describe both the receiving and scattering properties of the antenna when a plane wave is incident. In order to specify the properties of an antenna completely, it is necessary to consider the three operating conditions of the antenna illustrated in Figure 32. The surface S is designated as the terminal surface. It can be shown [11, 13, 14] that the basic equation which describes the antenna's scattered electric field $\overline{E}^s$ in terms of the operating conditions of Figure 32 is

$$\overline{E}^s (Z_\ell) = \overline{E}^s (0) - \frac{Z_\ell \, I(0) \, \overline{E}^r}{Z_a + Z_\ell} , \qquad \text{(A-23)}$$

where

Z_a = antenna impedance,

$\overline{E}^r$ = radiated field per unit current excitation, and

Z_ℓ = load impedance, and the (0) notation indicates $Z_\ell = 0$.

This equation indicates that the field scattered by an antenna at a given frequency and direction is a function of both the load and the antenna impedance. Equation A-23 may be transformed to a more convenient form as follows. First, set $Z_\ell = Z_a^*$ (complex conjugate of the antenna impedance) and then solve for $\overline{E}^s$ (0):

$$\overline{E}^s (0) = \overline{E}^s (Z_a^*) + \frac{Z_a^* \, I \, (0) \, \overline{E}^r}{2 \, R_a} . \qquad \text{(A-24)}$$

The terminal receiving current is given by

$$I(Z_\ell) = \frac{-V}{Z_a + Z_\ell} , \qquad \text{(A-25)}$$

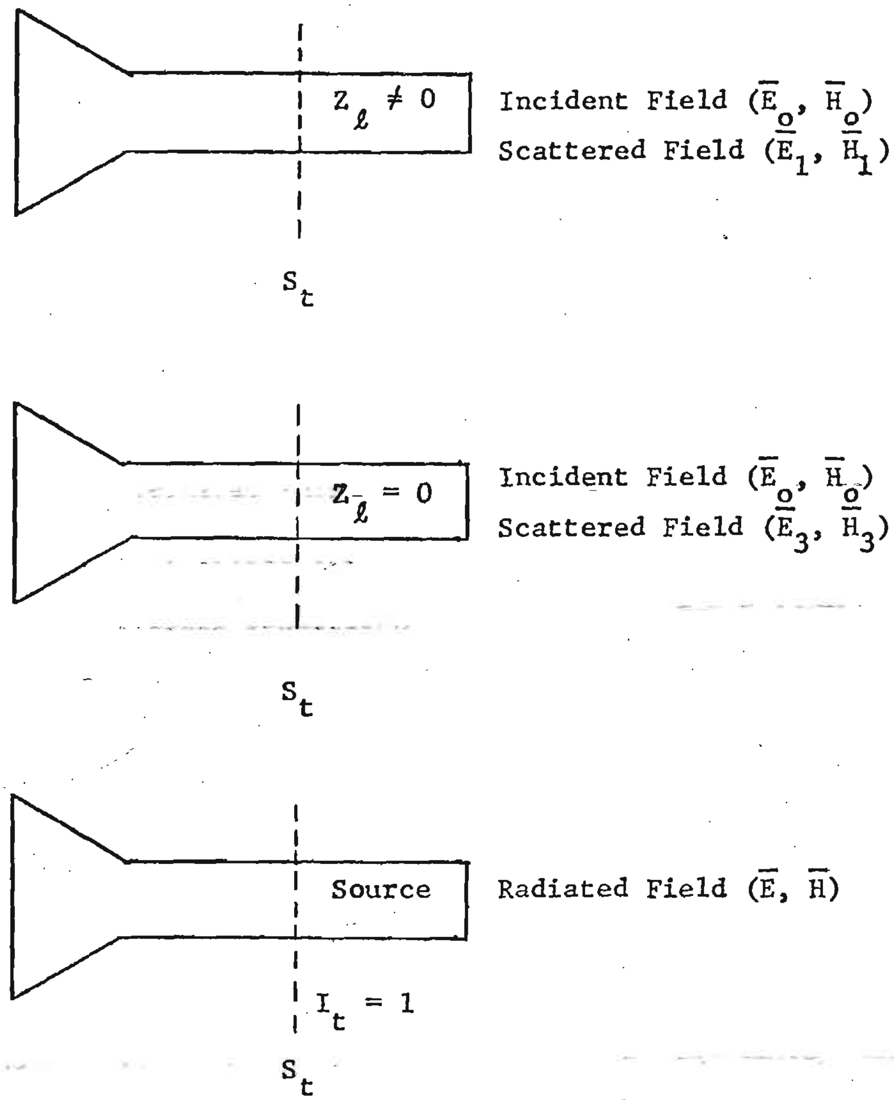

Figure 32. General antenna operating conditions

where V is the open circuit received voltage. Using Equation A-25 first with $Z_\ell = 0$ and then with $Z_\ell = Z_a^*$, and solving for I(0), one gets

$$I(0) = \frac{2R_a \, I \, (Z_a^*)}{Z_a} \quad . \tag{A-26}$$

Thus, Equation A-24 can be written:

$$\overline{E}^s \, (0) = \overline{E}^s \, (Z_a^*) + \frac{Z_a^*}{Z_a} \, I \, (Z_a^*) \, \overline{E}^r \quad . \tag{A-27}$$

Substituting Equations A-26 and A-27 into Equation A-23 results in the following expression for the scattered field:

$$\overline{E}^s \, (Z_\ell) = \overline{E}^s \, (Z_a^*) - \left[I \, (Z_a^*) \, \overline{E}^r \right] \Gamma_m \quad , \tag{A-28}$$

where

$$\Gamma_m = \frac{Z_\ell - Z_a^*}{Z_\ell + Z_a^*} \tag{A-29}$$

defines a modified voltage reflection coefficient. The first term of Equation A-28 may be interpreted as due to structural scattering and is a constant at a given frequency and direction; the second term is called the antenna mode scattering and is a function of the load impedance. The antenna mode scattering has the same pattern as the antenna when the antenna acts as a transmitter.

For most practical aperture type antennas, the maximum antenna mode scattering obtained for $\left| \Gamma_m \right| = 1$ is very much greater than the structural scattering in the "main beam" region. In short, the antenna is designed to be an efficient receiver of energy incident within the main beam. Thus assuming a short circuit at the antenna terminals $\left(Z_\ell = 0, \; \Gamma_m = -1 \right)$, the scattered field is

$$\overline{E}^s \, (0) = \overline{E}^s \, (Z_a^*) + I \, (Z_a^*) \, \overline{E}^r \quad , \tag{A-30}$$

which is approximated as

$$\overline{E}^s \, (0) = I \, (Z_a^*) \, \overline{E}^r \quad . \tag{A-31}$$

Recalling that $\overline{E}^r$ is the radiated field for a unit current excitation, one has

$$\overline{E}^r = -j \frac{\eta}{2\lambda} \overline{h}_t \frac{e^{-jkr}}{r} . \tag{A-32}$$

Also, the received current due to an incident electric field $\overline{E}^i$ (See Appendix III, Equations A-14 and A-15) is given by

$$I\,(Z_a^*) = \frac{\overline{h}_r \cdot \overline{E}^i}{Z_a + Z_a^*} = \frac{V_r}{Z_a + Z_a^*} . \tag{A-33}$$

Thus, the scattered field can be written as

$$\overline{E}^s\,(0) = \left(\frac{\overline{h}_r \cdot \overline{E}^i}{Z_a + Z_a^*}\right) \left(-j \frac{\eta}{2\lambda} \overline{h}_t \frac{e^{-jkr}}{r}\right) . \tag{A-34}$$

If we assume no polarization or pattern mismatch between $\overline{h}_r$, $\overline{E}^i$, and $\overline{h}_t$ for transmit and receive, as would be true for the antenna boresight axis with the same sense polarization, $h_r = h_t = h$, and Equation A-34 becomes

$$\overline{E}^s\,(0) = \left(-\frac{j\eta}{2\lambda}\right) \frac{h^2}{2R_a} \cdot \frac{e^{-jkr}}{r} \overline{E}^i . \tag{A-35}$$

This scattered field can be measured at a range r from the antenna. For antenna calibration, let a sphere be substituted for the antenna at range r and the scattered field of the sphere measured as

$$\overline{E}^s_{sphere} = \overline{E}^i \sqrt{\frac{\sigma_{sphere}}{4\pi}} \frac{e^{-jkr}}{r} . \tag{A-36}$$

The ratio of the antenna and the sphere scattered field is

$$\frac{\overline{E}^s\,(0)}{\overline{E}^s_{sphere}} = \left(\frac{-j\eta}{2\pi}\right) \frac{h^2}{2R_a} \sqrt{\frac{4\pi}{\sigma_{sphere}}} . \tag{A-37}$$

The factor $\sqrt{\frac{4\pi}{\sigma_{sphere}}}$, where σ = the scattering cross-section of the sphere, is known exactly for the sphere. Thus a measurement of the ratio of the scattered fields in magnitudes and phase together with a measurement of the antenna radiation resistance R_a allows the antenna height h to be determined. This measurement can be performed conveniently over a wide frequency band.

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13. ABSTRACT

This report presents the results of a program to study and investigate advanced measurement techniques for evaluating the performance of broadband antenna systems for use in high resolution radar systems. New techniques to measure, record, and analyze antenna gain and pattern performance were studied, and the feasibility of developing the necessary instrumentation to perform these measurements was investigated. Systems based on the use of broadband noise signal sources and systems using sweep frequency techniques were studied. It was concluded that systems using broadband noise signal sources would not be cost-effective. Preliminary design of a broadband amplitude-only electronically integrating sweep frequency system was completed. It was concluded that this system could be implemented immediately and that it would provide an effective first step in realization of the ultimate phase plus amplitude broadband antenna measurement system. The implementation and operation of sweep frequency amplitude plus phase systems were studied, and an effective approach to the realization of a phase/amplitude system was identified. Recommendations for implementing this approach are presented.

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